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AUTOMATED MARINE INTERNATIONAL
Newport Beach, California

Investigation
of
L-Band Shipboard Antennas
for
Maritime Satellite Applications

Final Report to:

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NATIONAL AERONAUTICS AND SPACE ADMINISTRATION

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SYNOPSIS

The concept of satellite services for the civilian maritime community has only just begun to receive serious attention by major governmental, industrial and technical organizations responsible for communications and navigation on the high seas. One of the elements needed by such organizations for planning purposes relative to the configuration of maritime satellite systems is parametric data on low-cost shipboard antennas which are practical for deployment by civilian maritime users. The International Telecommunications Union, at the 1971 World Administrative Radio Conference for Space Telecommunications, has just allocated 15 MHz of spectrum at L-band (1535-1542.5 MHz and 1636.5-1644 MHz) for maritime mobile services, on an exclusive basis.

Accordingly, the objective of this study was to perform and document a basic conceptual investigation of low-cost L-band antenna subsystems for shipboard use, by identifying the various pertinent design tradeoffs and related performance characteristics peculiar to the civilian maritime application, and by comparing alternate approaches for their simplicity and general suitability. The study was not directed at a single specific proposal, but is intended to be parametric in nature. Antenna system concepts were to be investigated for a range of gain of 3 to 18 dB, with a value of about 10 dB considered as a baseline reference. As the primary source of potential complexity in shipboard antennas, which have beamwidths less than hemispherical is the beam pointing or selecting mechanism, major emphasis was directed at this aspect.

Three categories of antenna system concepts were identified:

1. Mechanically pointed, single-beam antennas
2. Fixed antennas with switched-beams
3. Electronically-steered phased arrays

An analysis of general requirements and design constraints associated

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with the maritime environment established that the antennas should be simple, rugged and fairly small ($<1m^3$), that they should be circularly-polarized, and be configured to provide a coverage which is greater than hemispherical by an amount corresponding to negative elevation angles of 30° for larger ships (>150m in length) and as much as 45° for smaller ships (<150 m). It was also shown that ships rolling in heavy seas defines fairly high angular rates, and also produce changes in relative satellite bearing (azimuth) as well as elevation angles, which is shown to be a constraint on single-axis fan beam approaches.

Various types of beam pointing or selecting techniques were identified and analyzed, whereupon it was concluded that slaving the beam position to shipboard inertial references, after initial (manual) acquisition, can be most effective for the lower gain applications (<9 dB), and that automatic angle tracking of the satellite signal using a single-channel monopulse receiver scheme was most suitable for high gain antennas (>14 dB). In the moderate gain range of 9-14 dB, the performance and costs of the above two beam pointing techniques were shown to be comparable (within $300), when added installation costs of the slaved stabilization approach, for a gyrocompass repeater and a specially designed inclinometer, are included. The slaved stabilization technique requires periodic manual realignment by a crew member to compensate for the ships daily headway. This feature could constitute a reason for a lack of confidence in a worldwide maritime satellite network by the maritime and related commerce community, and is thus considered a disadvantage. Accordingly, the single-channel monopulse autotracking techniques is concluded as more desirable, pending compatability with the antenna concept.

A variety of antenna radiator designs within the three conceptual categories delineated above were analyzed for their suitability. Two system configurations were selected as especially credible candidates in the moderate gain range. The first is a mechanically pointed, single beam short-backfire antenna incorporating single-channel monopulse autotracking.

It can provide for a minimum gain of 13 dB in operation using a small (39 cm diameter by 10 cm deep) antenna radiator which should be very simple to produce.

The second configuration is a system of four linear 8-element arrays designed for switching of vertical fan beams (20° wide by 130° in elevation) about the periphery of the ship, with switching logic slaved to the shipboard gyrocompass and an inclinometer vertical reference. This quadruple Butler-fed array, with adjacent beam interpolation, would provide for a minimum net gain of about 9-10 dB in operation. While automatic angle tracking of the satellite signal is generally superior, the complexity of the receiver implementation in this array candidate is estimated to be relatively complex. Thus the slaved approach is considered more appropriate.

A comparision of the two selected candidates indicated that the mechanically-pointed single beam antenna should yield better performance at lower cost, assuming a single antenna subsystem installation. For ships severely constrained by overcrowding of equipment on masts and/or other suitable superstructures, and thus requiring dual antenna subsystems, the cost effectiveness superiority of the mechanically-pointed antenna is estimated to be reduced to a small, but still positive measure.

Thus, it is recommended that an L-band short backfire antenna subsystem, including a two-axis motor driven gimbal mount, and necessary single-channel monopulse tracking receiver portions be developed for demonstration of performance and subsystem simplicity. When integrated with the appropriate L-band transmitter, receiver, modems, etc., a prototype terminal would be available for system tests with experimental spacecraft.

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INVESTIGATION OF L-BAND ANTENNAS
FOR MARITIME SATELLITE APPLICATIONS

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INVESTIGATION OF L-BAND SHIPBOARD ANTENNAS FOR MARITIME SATELLITE APPLICATIONS

SECTION 1
INTRODUCTION

1.1 USER ANTENNAS FOR MARITIME SATELLITE SERVICES

It may be anticipated that the next decade will see a virtual explosion in the commercial exploitation of the last 12 year's developments in space technology. At the forefront will be the communications applications. In addition to the increased use of the existing INTELSAT relays for intercontinental point-to-point telephony, many other satellite communication systems are being planned. The well-known commercial applications include various forms of domestic television relay satellites and international satellites dedicated to mobile communications.

1.1.1 The Maritime Satellite Application

The "mobile" user category is specifically unique due to the impracticality of large antennas or otherwise complex user terminals, and due to their large population. For this category, the system design and cost-effectiveness trade-offs clearly show that user terminals should be as simple as possible, with the burdens of complexity placed in the system's space-segment. Within the mobile user category, the aeronautical application has received a good deal of attention in the past few years. The most pressing need here is for air traffic control in transoceanic flights. The maritime mobile application is just beginning to receive serious consideration, and has a number of unique characteristics.

First of all, the maritime mobile application incorporates the largest base of independent user terminals yet to be considered for any single satellite communication application; civilian or military. There are over 50,000 ocean-going vessels over 100 gross tons in size which are potential users. Within this decade, there is forecast to be almost 14,000 vessels over 1000 gross-tons, with almost 11,000 of these in the tanker, ore and bulk, and cargo/container ship categories of size greater than 10,000 gross tons. By contrast, there are forecast to be less than 1000 commercial international aircraft (worldwide) in 1980 which would be potential users of an oceanic aeronautical satellite system. Secondly, the pressing maritime service need is for communications. Recognizing the economic necessity of increased centralized management, and having exercised the half-century old CW telegraphy (Morse code) system to the fullest extent possible, the telecommunication departments of many shipping companies are becoming convinced that the time has come for a completely new system, i.e., satellite relay.

Ultimately, of course, it can be expected that a multi-functioned synchronous orbit satellite system which provides communications and navigation services will be required.[1] In contrast to the aeronautical application, however, where relative position <u>surveillance</u> (which may be considered the "inverse" of navigation) is critical to a truly active air traffic control system, the maritime user desires a direct self-navigation capability.

The need for reliable, high quality communications is primarily economic. The revenue-producing capability of the larger oil tankers is as high as $25,000 to $30,000 per day. Shipping companies seek to maximize vessel time at sea, as days wasted in ports waiting for replacement parts, personnel, or other administrative details can clearly be costly. Currently 93% of all maritime high seas communications, by message count, are via CW telegraphy (Morse code) using the limited HF and MF circuits.[2] These channels generally have poor reliability due to interference and propagation anomalies, which can cause long outages, and are also limited in

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(1) Dorian, Charles, "Application of Space Communications to the Maritime Mobile Service," Telecommunications Journal, Vol. 38-V, 1971.

(2) Heckert, G. P. and Mendoza, B. A., "Communication Satellites: A Management and Safety Aid for the Maritime Transportation Industry," IEEE National Telemetering Conference, Washington, D.C. 14 April, 1971.

capacity. The average delay experienced between transmission of a message to and receipt from a ship is about 12 hours. Thus, improved communications can be shown to be a key to increased efficiency. It is for this reason that ship size is an important parameter in estimating a user base for a maritime satellite system. It can be expected that the larger ships will be the first to equip with the necessary terminals. In addition to the efficiency-related communications such as diversion and position reports, estimated time of arrival data, bunkers, repair and storm damage reports, port services, cargo manifests, administration, weather and navigational advisory broadcasts, etc., there are other important needs for reliable communications such as distress and search and rescue (SAR) functions and crew (personal) communications.

Another unique feature of the maritime application is the capacity required. While the number of potential users is very large and the communication message categories are many, the per ship usage may be relatively low. Current traffic predictions vary over a broad range. The equivalent of eight to ten voice channels per ocean is predicted to be viable in an initial system, with tolerable delays. A realistic estimate for a follow-on generation system in time would be closer to 150 channels, worldwide. This is a small percentage of the capacity of the recently launched INTELSAT IV satellite. However, the latter's capacity is predicated on the use of very large, high gain ground terminal antennas, e.g., 29.5 meter (97 feet) diameter parabolas. Thus, satellite system sizing is not directly comparable. Figure 1-1 presents a simplified illustration of the maritime satellite system concept. Yet another unique feature is the potential use of different RF bands for ship satellite and satellite coast links. This feature does not effect the requirements for the shipboard antenna, except to that degree of satellite power conserved with cross-strapping of RF bands.

1.1.2 The Shipboard Antenna Problem

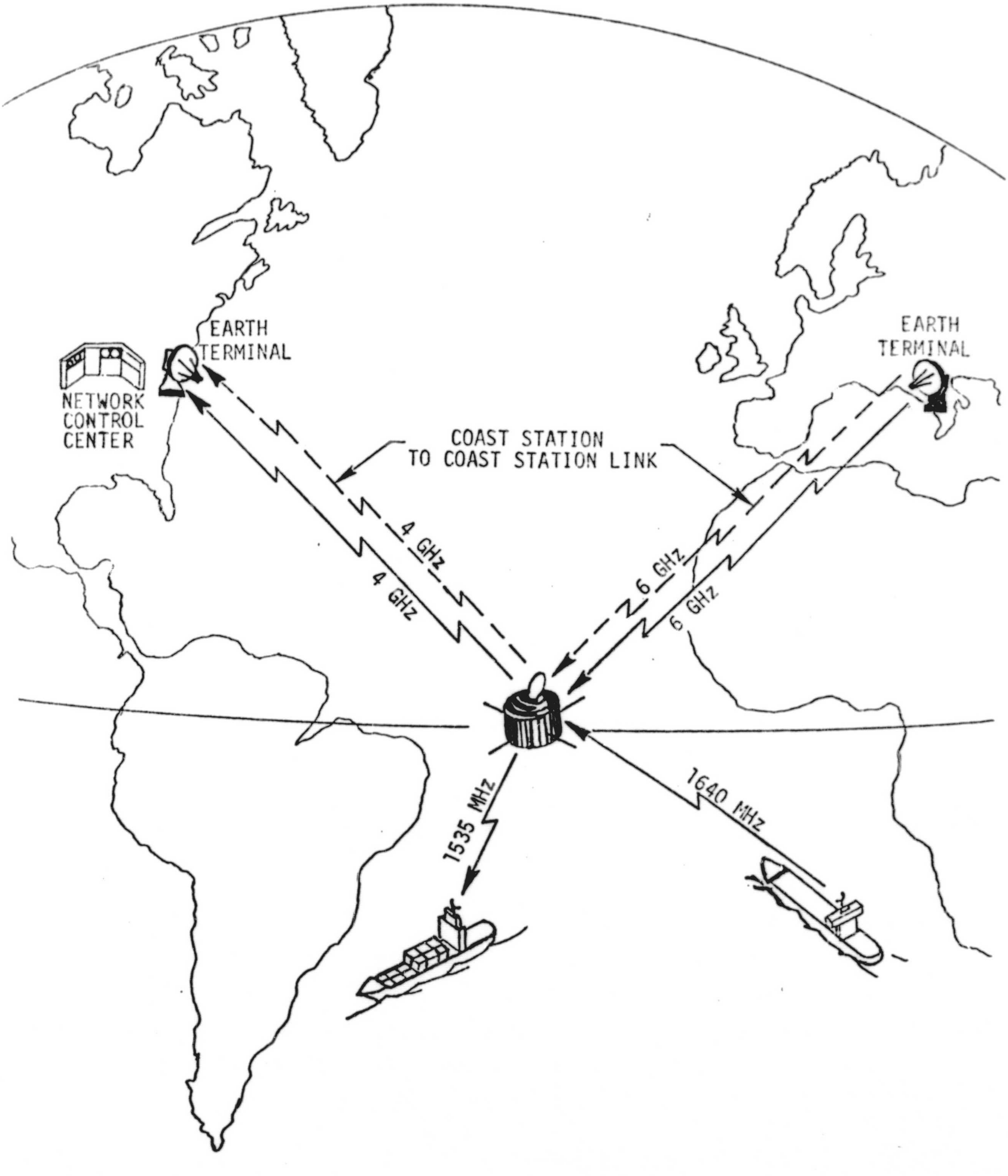

Figure 1-1 MARITIME SATELLITE CONCEPT

As noted, with such a large potential user terminal population, the cost effectiveness trade-offs for a maritime satellite system dictate simple, relatively modest antennas for the ships. Clearly, the simplest antenna would be a dipole. Of course, a dipole has negative gain in its longitudinal (end-fire) direction and the ship's antenna pattern must at least be hemispherical. That is, the worldwide satellite system requires hemispherical coverage of the ship antenna. While a hemispherical antenna with constant gain (0 to 1 dB) is not trivial to design, it would be quite simple in use in that no mechanism need be provided for pointing of the antenna beam. While theoretically possible as a system solution, a practical spacecraft design would be severely limited in its communications capacity. Relative to a simple hemispherical pattern, a ship antenna gain of only 3 to 4 dB would effectively double the capacity of the satellite system. A gain of 10 dB would increase the capacity of the same basic spacecraft design by an order of magnitude, etc.

Thus, there is strong motivation in the planning for a maritime mobile services satellite system to prescribe a certain amount of gain for shipboard terminals, even at the expense of shipboard equipment complexity. This is because power in space can cost $20-$30,000 per watt.

To date, operational-type satellite communication terminals for ships have been developed only for military use. Because of ships motion, and the relatively narrow-beam (less than 2 degrees) antennas employed, such terminals are very complex, and though they utilize small antennas (1.2 to 2.4 meter diameter), they typically require precision three-axis pedestals and sophisticated angle tracking subsystems. Such terminals are clearly not economically practical for the civilian maritime industry, and it is the basic objective of this study to investigate the feasibility of shipboard antenna subsystems which have some gain, e.g., 10 dB, but would be at least an order of magnitude less costly than the military-type noted above. For example, an antenna beamwidth of 2 degrees corresponds to a gain of about 38 dB, while a gain of 10 dB implies a beamwidth of over 50 degrees. However, it is not a priori clear that such a large increase in beamwidth, or corresponding beam pointing tolerance, permits order-of-magnitude reductions in control system complexities.

Within the low to moderate range of gains considered, a wide variety of antenna subsystem concepts are possible, including single beam antennas which are mechanically pointed, switching between multiple independent, fixed antennas, or electronically scanned arrays. The identification of the trade-offs between these concepts, as well as between various radiating elements and basic pointing techniques, specifically for the maritime satellite application, are the subjects of this study. In general, such data related to the feasibility of using simple, moderate gain shipboard antenna subsystem on commercial maritime vessels is currently needed for planning purposes relative to maritime satellite services.

1.2 STUDY OBJECTIVES AND SCOPE

1.2.1 Principal Goals

The fundamental objective of this study is to perform and document a basic conceptual investigation of L-band antenna subsystems of moderate gain (3-18 dB) for shipboard use, by identifying the various pertinent design trade-offs and related performance characteristics peculiar to the commercial maritime application, and comparing alternative approaches for their simplicity and general suitability. The data is intended for use by system planners in postulating maritime satellite system configurations, link sizing and economic trade-offs.

It is not the specific intention to propose a single antenna subsystem design approach as optimum for all ship categories. In point of fact, it might be expected that a fully operational future satellite communications system will involve the use of different shipboard antenna configurations, depending on ship category and desired quality, etc. Just as exists in the HF and VHF bands today, as well as other applications, it would be normal that different commercial electronics equipment suppliers market somewhat different antenna designs, all of approximately equivalent gain, but with variations in cost and quality, etc.

As noted in the preceding section, parametric data on the practicality of achieving or requiring a certain amount of user antenna gain is required for planning purposes. For example, specifying higher gains for user terminal antennas would permit the use of a larger capacity, i.e., number of channels, with a spacecraft of a given transmitted power capability. As the cost of the entire space segment is essentially directly related to satellite EIRP, the per channel usage costs of the satellite network would be less if all ships employ higher gain antennas. This, of course, must be balanced against the ship owners' investment in the shipboard terminal equipment. As the number of potential ship users is very large, this economic trade-off generally results in the need to minimize shipboard equipment complexity. However, there are both practical and economic limits to launch vehicle capability and satellite power, and thus several previous AMI studies[3][4] concluded that user antenna gains on the order of 10 dB are needed for a credible system plan. More generally, gains from 3 dB to 18 dB are of interest, reflecting different satellite design concepts and system capacities. Accordingly, this study addresses antenna concepts for gains in this range, with a nominal 10 dB as a baseline reference value.

In addition to the economic trade-offs, the parametric data on antennas is required for a host of other system considerations, including channel frequency spacing. The initial system may be considerably power limited in the satellite-to-shipboard link, and as such could be capacity limited to the order of eight channels (of voice or equivalent capacity each). In such a system, FDMA (frequency division multiple access), with independent signal carrier accessing a single satellite transmitter power amplifier, will probably be used for system flexibility, wherein as much as 2.6 dB of link power savings can be realized if intermodulation-controlled channel spacing is used. This could correspond to a factor of 80% in capacity. (More specifically, it represents approximately the difference between a 5 or 6 channel capacity and an 8 or 9 channel capacity, respectively.)[3]

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(3) "A Study of Maritime Mobile Satellite Service Requirements, Frequency Planning, Modulation and Interference Analysis," in 6 Volumes, U.S. Coast Guard Contract DOT-CG-00505A, by Automated Marine International, Newport Beach, California. 15 October 1970.

(4) "Maritime Services Satellite - System Definition Study." U.S. Department of Transportation - Transportation Systems Center. Contract DOT-TSC-98, by Automated Marine International, Newport Beach, California. 31 May 1971.

Such channel spacing, however, requires considerably more RF spectrum than the simple side-by-side stacking of channels (e.g., by a factor of about 4.4 for an eight channel capacity). The viability of an initial maritime system may depend on I/M* controlled spacing, and since satellite power can be traded for shipboard user antenna gain, a relationship exists between shipboard antenna complexity, system capacity and L-band spectrum requirements. Such channel spacing should not be considered unrealistic for an initial system design. The initial system will not require the same capacity that the second and third generation will need; thus, a given RF spectrum allocation, such as 7.5 MHz for example, could be effectively used with an initial satellite system providing a small number of channels (e.g., 10 per ocean) with maximum power efficiency and then re-used in second and third generation satellites, of considerably increased EIRP capability, to provide the then required capacity (e.g., 150 channels worldwide in time) using adjacent channel spacing.

Thus, the practicality of achieving a certain shipboard antenna gain is a major factor in the consideration of the subject satellite system design. While there are a wide variety of antenna radiator configurations which are possible and interesting, most are relatively inexpensive in volume production, and the primary source of potential complexity lies in the provision for pointing of the directive beam. It is this facet of the shipboard antenna subsystem which is most significant in terms of uncertainty and potential cost. The fact that the antenna is to have some positive value of gain implies a beamwidth less than that of a hemispherical pattern, and thus some mechanism must be provided for the pointing of a movable beam or for selecting the proper beam from a set of differently oriented fixed beams.

The classic method of beam pointing is that of automatic angle tracking of the received satellite signal's direction, implemented with either a rotating feed or set of feeds such that a comparison of multiple beams can be used to generate a pattern null in the signal (boresight) direction. Off-boresight angular positions then produce a signal voltage which is used,

*I/M - intermodulation

after certain processing, to drive beam position controllers (e.g., antenna drive motors). While this approach has, of course, numerous merits, including accuracy, it would be desirable that an alternative scheme be shown suitable, or at least a version of the approach which is significantly less complex than the traditional monopulse and conical-scan angle tracking systems used with narrow-beam antennas. Thus, the most fundamental objective of this study is the evaluation of the relative complexities and merits of various beam pointing techniques for an L-band shipboard antenna of modest gain, e.g., 3-18 dB.

1.2.2 Scope

The scope of this investigation of L-band ship antennas for satellite communication services is limited to theoretical analyses and review of pertinent results of related efforts. The contract Statement of Work defines that the study include[5]

a. A parametric engineering analysis of shipboard antenna tracking concepts operating in the 1600 MHz frequency band.
b. A trade-off of relative performance and cost and related conceptual engineering design analysis of at least two shipboard antennas having gains of 10 dB (such as switched multiple beams and narrow beam antennas having peak gains of between 3-18 dB).
c. The consideration of L-band antenna tracking schemes with mechanical drive freedom, switched multiple antenna beams, and mechanization of manual pointing of narrow beam antennas based on an operator observing a communication receiver's AGC level.
d. A general assessment of the feasibility of at least two L-band antennas for shipboard use with satellites with emphasis on design complexity, cost and performance.

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[5] National Aeronautics and Space Administration, Contract NASw-2165. NASA Headquarters, Control No. 10-9612. March, 1971.

1.3 APPROACH

1.3.1 The Maritime Satellite Service Requirement

The first in-depth technical study of maritime mobile satellite services was performed last year by Automated Marine International for the U.S. Government.[3] The study was managed by the U.S. Coast Guard and co-sponsored by the Department of Transportation's Office of Telecommunications, the U.S. Office of Telecommunications Policy and the Department of Commerce's Maritime Administration. Its six volumes addressed the potential user population size, the service and capacity requirements, signal modulation and access, frequency planning and interference anaylsis and related system performance trade-offs.

The primary purpose of this study was to provide technical data in support of U.S. planning for this year's (1971) World Administrative Radio Conference (WARC) held by the International Telecommunications Union in Geneva. Until this conference, no frequencies were allocated to maritime mobile for space services, and one on the functions of the 1971 WARC was to make spectrum allocations for such services. As WARCs are held only once every 8 or 10 years, it was considered vital that maritime requirements be well established. One of the conclusions of the aforementioned study, hereafter referred to as the 1970 DOT-CG Study, was related to the RF frequencies most suitable for the subject application. As with aeronautical services, two bands had generally been considered for use - the existing maritime mobile line-of-sight VHF band in the 156 MHz to 174 MHz region and the mobile UHF band in the 1540 MHz to 1660 MHz region.

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[3] 1970 DOT-CG Study. Op. Cit.

In strong contrast to the aeronautical mobile VHF band in the 118 MHz to 136 MHz region, the existing VHF maritime mobile band is shared with a wide assortment of other services, primarily land mobile and fixed, but also including broadcast and television in some non-U.S. countries.

Frequency Planning

In August of 1968, the U.S. Federal Communications Commission (FCC) opened Docket 18294 inquiring into matters pertaining to the 1971 WARC, the second such conference to consider space telecommunications. The notices of inquiry, which followed, were primarily intended to solicit non-government views relating to frequency requirements of the several generic radio services. Government needs were accommodated through a separate agency called the Inter-Governmental Radio Advisory Committee (IRAC). All views are then coordinated in joint session through IRAC, FCC, the Presidents' Office of Telecommunications Policy (OTP) and the U.S. State Department.

By August of 1969, the FCC had issued five notices of inquiry in subject docket, had received and commented upon hundreds of industry and user responses, and had subsequently issued via the State Department, the Preliminary U.S. Views to the 1971 Space WARC. The preliminary views with respect to maritime were to recommend space service authorization of the following frequencies:

VHF: 156.4-157.4 MHz and 173-173 MHz
UHF (L-Band): 1535-1537.5 MHz and 1657.5-.660 MHz

At the same time, it was proposed that aeronautical mobile receive for its service needs at L-band the following:

UHF: 1537.5-1557.5 MHz and 1637.5-1657.5 MHz

Thus, maritime was to receive two allocations of 2.5 MHz at L-band separated by 122.5 MHz. Aviation was to receive two 20 MHz bands separated by 100 HHz.

The results of the 1970 DOT-CG Study showed rather conclusively that the Preliminary U.S. Views to WARC-ST were unsuitable for the following reasons:

a) The maritime mobile space service was forecast by 1980 to need a minimum of 5 MHz of spectrum in each transmission direction.

b) The VHF band 156.4-157.4 MHz was heavily congested with not only maritime mobile users but also land mobile and fixed services. Many of the latter two groups did not conform to maritime channel spacing or emission standards, and shared space/terrestrial services would not be feasible due to interference. Further, the band 173-174 MHz was used for TV broadcasting in France and Japan and it could neither be shared with maritime nor cleared for maritime. Cleared channels were mandatory at VHF and clearing enough spectrum and assuring same appeared to be a remote possibility.

c) The L-band frequencies were potentially interference free, but the 2.5 MHz of spectrum was grossly inadequate for future maritime needs, and aviation and marine should be authorized a common up-and down-link frequency separation (for possible joint satellite use). Further, for maritime to avail itself of a minimum cost spacecraft with hemispheric (north and south) coverage, it would need to utilize a shipboard antenna with a gain of about 10 dB, the economic practicality of which was not established. (This compares with a 0 dB antenna at VHF with the same spacecraft sizing and capacity.)

The study also concluded that as an alternative to both VHF and L-band, the region around 400 MHz is a near-optimum in terms of satellite power minimization for a given capacity and 0 dB gain (hemispherical) user antennas. This conclusion was also made by independent studies in England; however, the unsuitability of the 400 MHz region due to sheer unavailability of spectrum is so pronounced that little support was given for such proposals at the WARC. While the final proposals to the WARC included a request for a modest amount of spectrum in the region near 400 MHz, only 100 kHz was allotted, at 406.0 to 406.1 MHz, for earth-to-space EPIRB's (Emergency Position Indicating Radio Beacons), and this allocation is to be on a shared basis.

At VHF, the WARC authorized maritime use of space for safety and distress purposes on an exclusive basis in two bands of only 100 kHz each (157.3125 to 157.4125 MHz and 161.9125 to 162.0125 MHz). Even at that, use of these allocations is not permitted until 1976, and it was resolved that the 1974 Maritime ITU conferences should consider if and to what extent space techniques should be introduced in this band.

At L-band, the results of the 1971 WARC-ST were far more fortuitous for the maritime mobile application, considering that as late as 1969 virtually no organized effort was made by any civilian maritime entity to address the subject. A full 7.5 MHz was allocated in each transmission direction to maritime for satellite use on an exclusive basis, and an additional 1 MHz in each direction shared with aeronautical mobile. The specific authorization is:

1,535 MHz - 1,542.5 MHz	Maritime Mobile - Satellite (Space-to-Earth)
1,636.5 MHz - 1,644 MHz	Maritime Mobile - Satellite (Earth-to-Space)

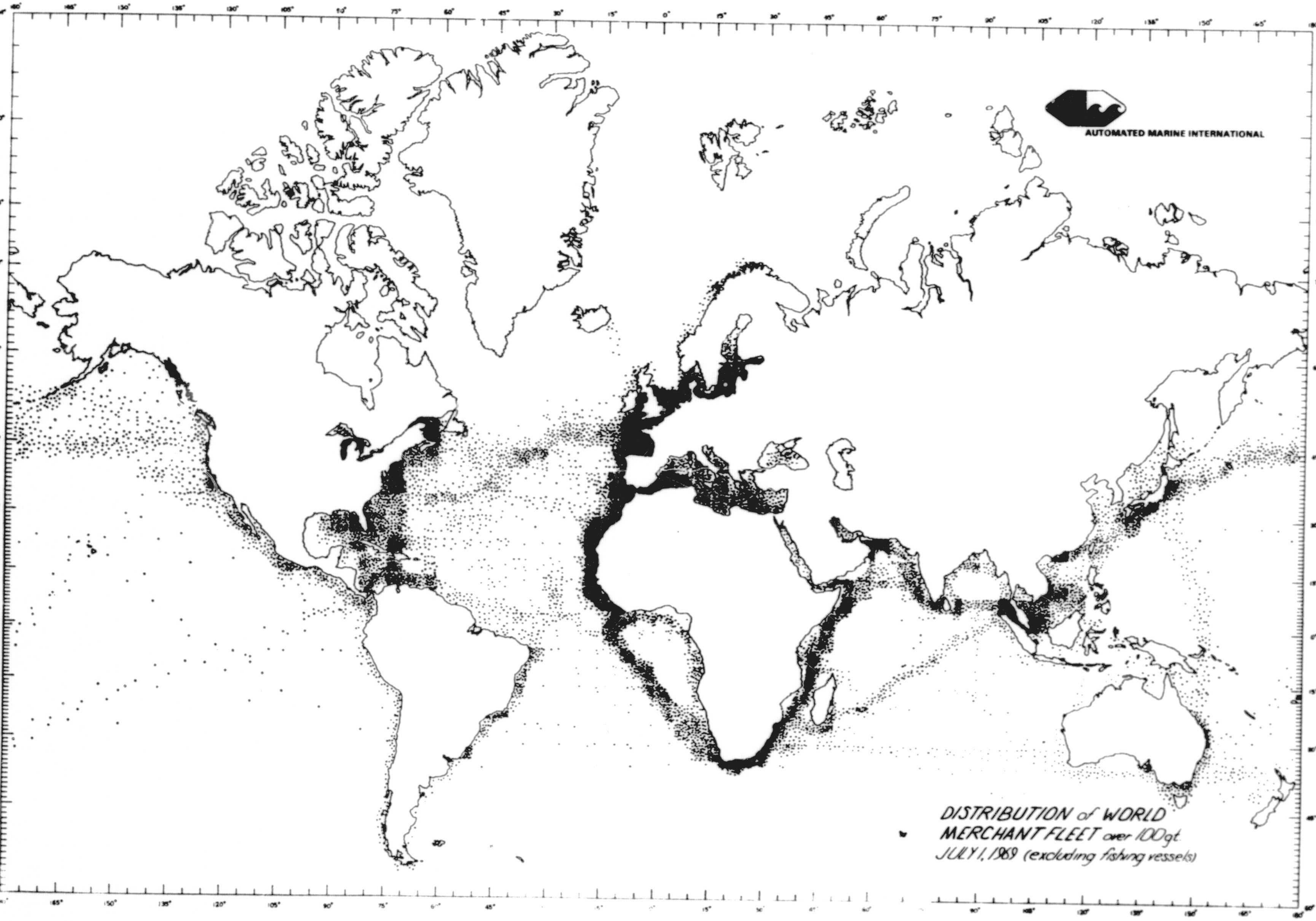
1,542.5 MHz - 1,543.5 MHz	Aeronautical Mobile (R) - Satellite, Maritime Mobile - Satellite
1,644 MHz - 1,645 MHz	Aeronautical Mobile (R) - Satellite, Maritime Mobile - Satellite

User Population and Distribution

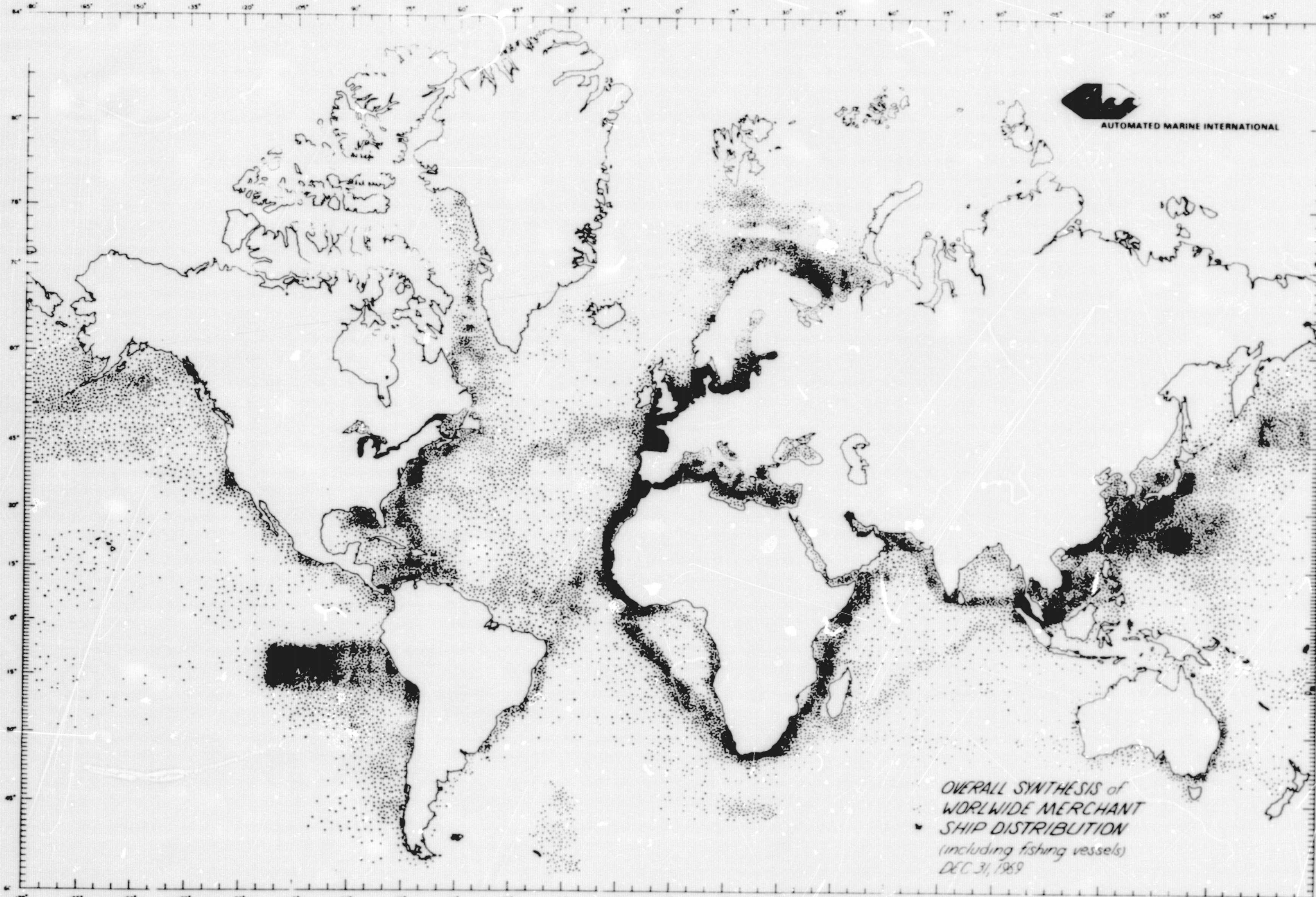
In addition to the study of potential interference between the civilian maritime and other services, the 1970 DOT-CG study attempted to ascertain the number and type of channels required, i.e., the amount of spectrum needed and related signal characteristics. In order to do this, it undertook in-depth surveys of the number of potential user ships in the world, and their percentage of time at sea. Figures 1-2 and 1-3 show the concentrations and locations of all ships at sea on a specific day in 1969, as typical, with and without fishing vessels included. Each dot represents a ship (over 100 gross tons), based on actual statistics derived from plotting the routes corresponding to all arrival and departure announcements published daily in Lloyd's List. Figure 1-4 shows the predicted total distribution for 1980, based on various growth analyses.[3] Table 1-1 indicates the number of vessels at sea projected in 1980, by category and size.

While the total number of ships in the world was predicted to grow only slightly (from 50,276 to 54,114 vessels over 100 gross tons), the percentage at sea at a given time is predicted to increase significantly due to the increased efficiency of new (and replacement) ships. The worldwide distribution of the potential users clearly indicates the necessity for worldwide coverage by the satellites. An example coverage plot available with a three-satellite system is shown in Figure 1-5.

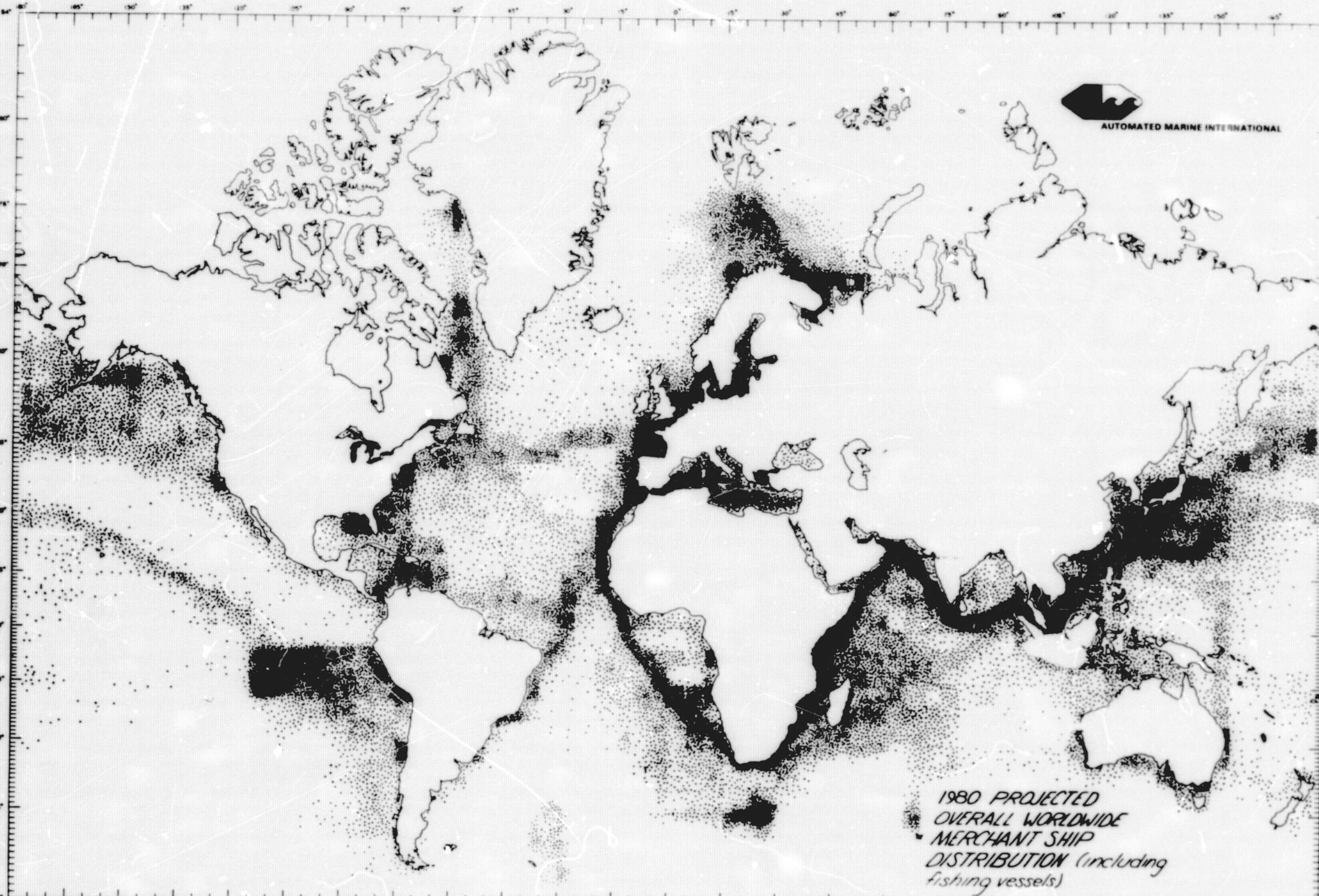
(3) 1970 DOT-CG Study. Op. Cit., Vol. I.



DISTRIBUTION of WORLD MERCHANT FLEET over 100gt.
JULY 1, 1969 (excluding fishing vessels)



OVERALL SYNTHESIS of
WORLWIDE MERCHANT
SHIP DISTRIBUTION
(including fishing vessels)
DEC. 31, 1969



1980 PROJECTED OVERALL WORLDWIDE MERCHANT SHIP DISTRIBUTION (including fishing vessels)

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TABLE 1-1
VESSELS AT SEA WORLDWIDE (1980 FORECAST)*
(Total Number of Vessels Shown in Parentheses)

Size, Gross Tons	Tankers	Ore/Bulk	Cargo/ Psgr.	Fishing Vessels	Fishing Fac- tories	Other	Total
≥100	5559 (7039)	3519 (5370)	4935 (19565)	14400 (18455)	546 (701)	254 (2984)	29213 (54114)
≥1000	4455 (5426)	3519 (5370)	4095 (11160)	1190 (1400)	420 (512)	200 (1339)	13879 (25207)
≥6000	4072 (4848)	3519 (5370)	3105 (6416)	- -	193 (227)	65 (252)	10954 (17013)
≥10000	3938 (4633)	3235 (4900)	1320 (1908)	- -	108 (127)	25 (65)	8626 (11433)

* Includes vessels in Great Lakes, Mediterranean, etc. which may not use satellites.

Communication Service and Capacity Requirements

A second study for the Department of Transportation has just been completed. This was a smaller effort addressing the operating doctrines of a Maritime Satellite (MARSAT) Services System, network configurations, and spacecraft and user terminal subsystem configuration trade-offs.[(4)] This study was performed for the Telecommunications Division of DOT's Transportation Systems Center (TSC). This 1971 DOT-TSC Study specifically excluded the trade-offs associated with shipboard user antenna subsystems. The user antenna subsystem was isolated as an item for a separate study (contained herein), as it is considered an element of critical technology in relation to the viability of and the planning for a Maritime Satellite System.

Both the 1970 DOT-CG Study and the 1971 DOT-TSC Study addressed the

(4) 1971 DOT-TSC Study. Op. Cit.

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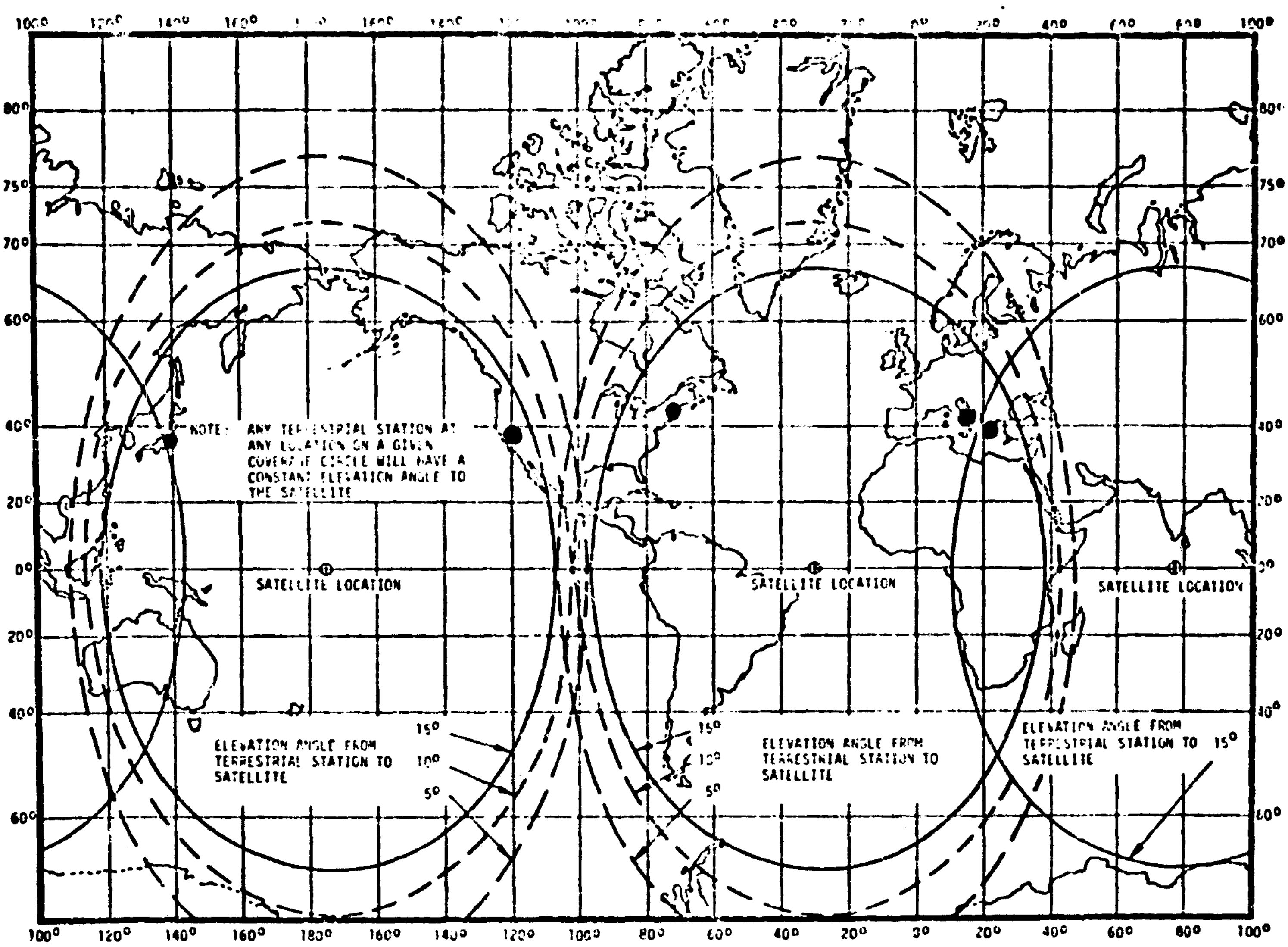

MARITIME SATELLITE EARTH COVERAGE
CONTOURS FOR LOW ELEVATION ANGLES

FIGURE 1-5

communication service requirements in detail. It was shown that most of today's maritime communications is message traffic using CW telegraphy (Morse Code) at HF. The average message delay in each direction is 6 hours. A careful survey of potential users (various worldwide shipping companies) has indicated a requirement for much more capacity, both for data (hard copy telex and graphics) and radiotelephone. Using the year 1980 as a point of reference in terms of operational capacity requirements in an initial Maritime Satellite system, it was estimated, on the basis of user surveys, that about 4500 ships would be equipped for satellite service. Equipage would start slowly in 1975, with the very large vessels - tankers, ore and bulk carriers and fast loading/unloading cargo vessels. This user base includes one out of two of the tankers in excess of 10,000 gross tons (gt), (2200) and one out of four of the ore and bulk carriers over 10,000 gt (1200) and one out of two of the container-type vessels over 10,000 gt (380) and one out of four of the general cargo and passenger ships over 10,000 gt (300) and one out of five of the fishing vessels over 1000 gt (370).

The satellite system capacity required for this user base was estimated to be the equivalent of 8-10 duplex voice channels in the busiest two oceans. This result is based on careful analysis of telex word counts according to ship category, the required use of telephone for ships business and crew (personal) communications, etc. and is a strong function of the amount of delay assumed acceptable in establishing a particular link. While essentially instantaneous access by ships would be provided for emergencies, the above capacity prediction is based on an average delay on the order of 10 minutes being acceptable for routine communications.

In addition to the capacity requirements analysis, the 1970 DOT-CG Study considered in detail the power and bandwidth trade-offs associated with various modulation techniques. As the required radiotelephone service is that which effectively sizes the system, the analysis of voice modulation schemes was emphasized. For a given satellite/launch vehicle category, the combination of the required system capacity and individual channel power requirements leads directly to required shipboard gain.

Figure 1-6[3] shows a comparison of representative voice signal modulation techniques, in terms of required carrier-to-noise density, C/N_0, at the receiver for a given articulation index (a measure of intelligibility).

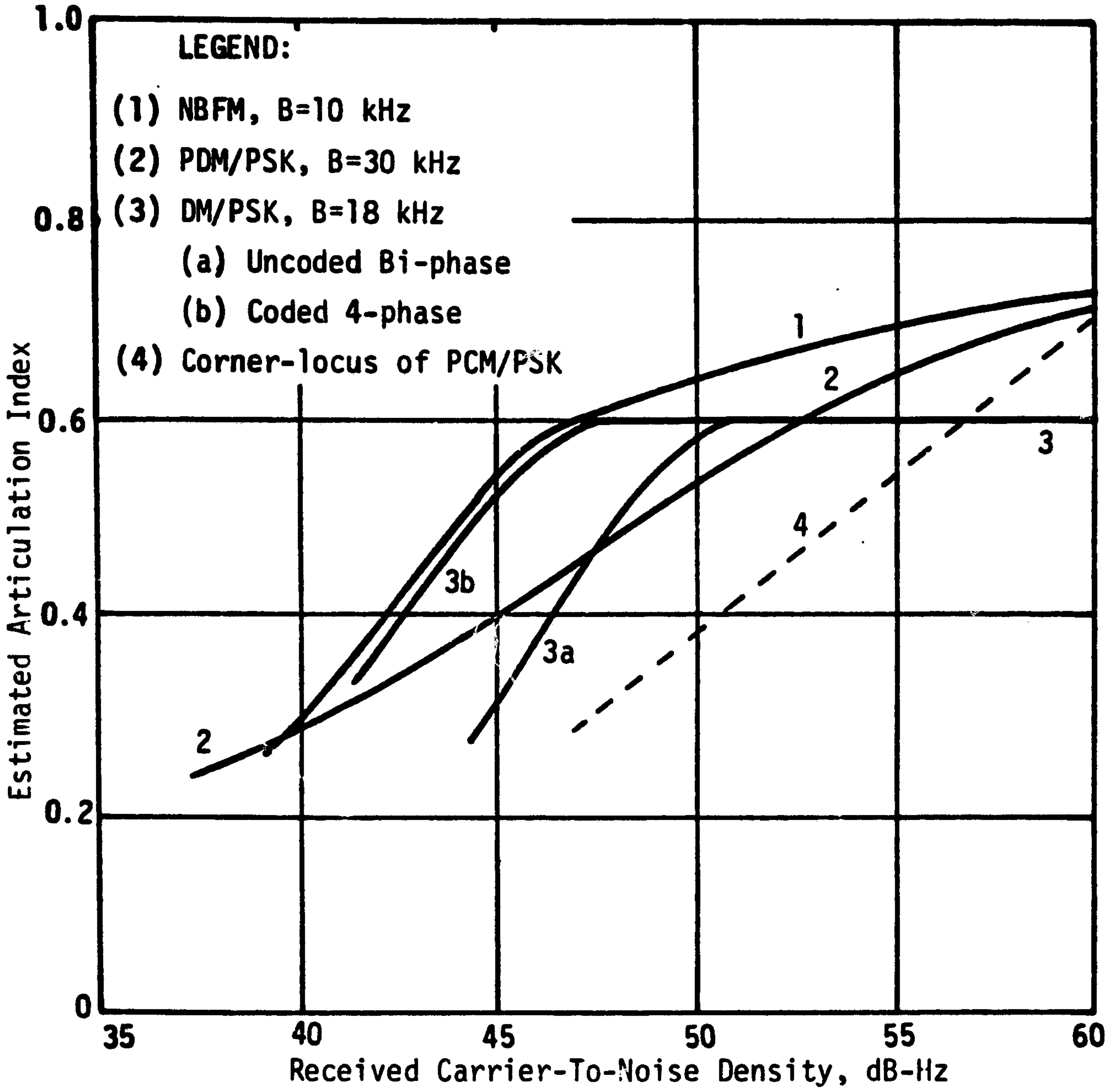

Figure 1-6 APPROXIMATE COMPARISON OF SELECTED VOICE MODULATION AND PROCESSING SCHEMES IN TERMS OF RELATIVE ARTICULATION INDEX

Generally speaking, NBFM (narrow-band frequency modulation, with clipping and phase-locked demodulation), PDM/PSK (pulse-duration modulation of the baseband with phase-shift keyed modulation at RF), and DM/PSK (delta-modulation encoding with PSK at RF) all exhibit good power efficiency.

A value of C/N_0 of 45 dB-Hz is representative of that which should be provided at the ship receiver for each satellite channel, the large majority of the time. When the spacecraft is constrained to a Thor-Delta launch and a single earth-coverage antenna, with an 8 to 10 channel capacity, the result of the above requirement for received C/N_0 is effectively the specification of shipboard antenna gain at about 10 dB.

Navigation, Surveillance and Traffic Advisory Services

The maritime community, on the high seas, navigates celestially, and in many areas they navigate by radio beacons, radio direction finder stations, Loran A and/or C, by Decca Navigator and by Consol. Omega and Transit based navigation are not in widespread use but can be expected to increase over the next 10 years. Loran A is the most widely used high seas system where it is available (almost all of the Northern Hemisphere only), and yields a 2-sigma position accuracy of about 2 n mi. Position reports are transmitted to home offices daily or every other day.

The 1970 DOT-CG study was constrained to use of the DOT-National Plan for Navigation (NPN)[6] which did not indicate a requirement for

(3) 1970 DOT-CG Study, Volume III. Op. Cit.

(6) Department of Transportation - National Plan for Navigation. May 1970.

development of a new maritime navigational or surveillance system (i.e., satellites). This plan states that "U.S. commercial maritime interests would be 90% satisfied with complete coverage of the North Atlantic and North Pacific...(and) a high seas position fix accuracy (now and for the next 20 years) averaging 2 n mi. rms, with an upper limit on system error of 5 n mi. ...and a fix interval not exceeding 2 hours." (It should be noted that this requirement is for U.S. essential trade routes only and would not necessarily be applicable for worldwide shipping.) For coastal and confluence areas, the DOT-NPN states, "the mariner would be satisfied with a radio navigation system having the capability of fixing position (continuously) to a repeatable rms accuracy of 1/4 n mi. out to a range of 50 n mi. off the coast. ...Recognizing, however, that a large number of the users of the coastal/confluence zone are unable or unwilling to pay the price for this capability, the system should provide service of useful quality within the economic reach of all mariners, such that no boat or ship need be lost nor endangered because of lack of knowledge of position."

Consistent with these DOT-NPN constraints on the study, it was adjusted that the navigation aids available today or planned in the near future (e.g., Omega) were adequate to meet the high seas requirements of the maritime community. Further, with the pending availability for civilian maritime use of single channel (400 MHz) TRANSIT (U.S. Navy Navigation Satellite) Systems,* it was considered questionable whether a new satellite system specifically intended for position determination could be justified.

However, in contrast to the DOT-NPN, the Inter-Governmental Maritime

* At prices generally quoted between $25-$30,000 complete with receiver, computer and display.

Consultive Organization (IMCO)[7] does consider an eventual maritime satellite requirement for the following services:

- "interrogation of the land-based station by a mobile craft station for obtaining position information possibly followed by environmental, meteorological and oceanographic information, or
  regular interrogation of the mobile craft station in appropriate time intervals by the land-based station and transmission of position information etc. to the mobile craft.
- initial position determination providing accuracies of the order of 1-2 n mi. As technology develops, and considering cost effectiveness, the accuracy may be improved to be available for navigation near coasts and in narrow passages.
- automatic warning of ships which are continuously tracked by the system, in cases of approaching shallow waters, underwater obstructions, drilling and production platforms, etc.
- advising ships which are continuously tracked by the system on anti-collision actions and on avoidance of continuously tracked navigational hazards, e.g., icebergs.
- automation of the position-reporting system based on position information as mentioned in the first item. Thus, the present repeated individual reporting actions on mobile craft could be abolished.
- traffic control including collision warning especially in converging areas subject to the radiodetermination system providing sufficient relative accuracy (see second item)."

Thus, IMCO clearly envisions ultimately a satellite based navigation, surveillance, position-reporting and traffic advisory system. They are quick to recognize, however, that there is no ordered and regulated traffic control in the maritime industry today and technology notwithstanding, adequate international regulatory agreements may not be forthcoming in this decade.

Thus, it would appear that the maritime community in general believes

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(7) Inter-Governmental Maritime Consultive Organization (IMCO), 7th Session of the Subcommittee on Radiocommunications, Report to the Maritime Safety Committee, July 1970.

it has adequate navigational aids and capability on the high seas, and that reliable satellite communications and position reporting could significantly reduce the search part of SAR. However, in coastal trade and in outer and inner confluence areas, there is increasing voluntary establishment and use of sea lanes. These lanes require relatively precise position fixing accuracy (e.g., 1/4 n mi., 1-sigma for coastal and outer confluence areas), and it is not known whether geostationary satellite based systems would be more cost-effective than terrestrially based systems. It is also not known whether fleet owners and operators would pay for the cost of any new navigation/surveillance system, although there are indications that if equipment costs (over and above that required for satellite communications) are on the order of $10-$20,000, and the system could be used for high seas position-fixing as well, then it probably would elicit adequate support.

Clearly, detailed trade-off studies are required to ascertain the benefit versus costs of a new positioning system for use both on the high seas and in outer confluence areas and whether such a system (i.e., satellites and launch vehicles) might be shared with aviation to reduce per user costs. Users may generally want to reduce on-board vessel equipment and complexity, enabled by shore-based computation of position and subsequent relay back to the ship. An ancillary high seas benefit of such a system is the potential elimination of the search phase of search and rescue (SAR), but such a benefit is difficult to assess in terms of dollars saved per year.

A major factor to be considered with geostationary satellite based positioning system is the attendant impositions, if any, on the shipboard antenna system. If the satellite system incorporates two or more satellites per ocean to effect a range-range positioning concept, as is usually considered best, and if narrow beam antennas are required for efficient system (power) sizing, then two independent antenna subsystems could be required on each ship.

1.3.2 Related Efforts

As indicated, the subject L-band antenna study is to be conceptual and parametric, with the primary intent being to identify the various pertinent design trade-offs and related performance parameters of shipboard user antenna subsystems peculiar to the commercial maritime application. No other previous investigation of commercial shipboard satcom antenna subsystems is known to exist, and even though various shipboard satcom antennas have been implemented and used, little analogy exists from which pertinent conclusions may be drawn.

Experimental satcom terminals have been placed on ships and used in demonstration tests with NASA's Applications Technology Satellites. Most of these have been between ATS-1 and -3 at VHF (135-150 MHz) and ships of the U.S. Coast Guard, NASA, a steamship company and certain foreign countries (England, Holland). The U.S. Coast Guard experiments were performed with the cutters USCGC Klamath, Staten Island, Glacier and Casco.(8) Grace Line's S.S. Santa Lucia also participated in tests. In these demonstrations, the antennas employed were very simple, and of course, at VHF. For example, the CGC Glacier's receiving antenna was a simple crossed-yagi (three crossed dipoles on a pole) providing 7 dB of gain at 135.6 MHz.(9) It was manually pointed using a simple home TV-

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(8) "Eighth Report by the International Telecommunications Union on Telecommunication and the Peaceful Uses of Outer Space." ITU, Geneva, 1969.

(9) Ware, J.N. "VHF Shipboard Tests on the U.S. Coast Guard Cutter Glacier (WAGB-4) - Final Report", U.S.C.G. Washington Radio Station, Alexandria, Virginia. 29 May 1968.

type rotator. A helical folded dipole type antenna with a hemispherical pattern was used for transmitting at 149.2 MHz, providing 0 dB of gain.

Limited tests were performed at L-band with the unintentionally spinning ATS-5 satellite from the Humble Oil Tanker S.S. Manhattan in the spring of 1970, and other very limited ATS-5 tests are being conducted in 1971. The antennas used on the S.S. Manhattan were simple parabolas mounted on adjustable tripods.[10] Two antennas were used, one of 0.91 meter (3 ft.) diameter with a cavity-backed spiral prime-focus feed and providing a gain of 21 dB, and the other of 0.61 meter (2 ft.) diameter and a gain of 17 dB. The simple deck-mounted tripods supporting the antennas were manually adjustable, and tests were performed by manually adjusting the antenna orientation (typically each morning) for maximum received signal level. The S.S. Manhattan is quite large, of course - over 100,000 dead weight tons and thus experiences relatively small roll and pitch dynamics. In any case, it is understood that the ATS-5 tests were primarily performed while the Manhattan was essentially motionless in ice, so the lack of any mechanism for automatically steering the antenna was not pertinent to the experiment. In all these non-military experiments, the shipboard terminals were configured for test purposes only, and in no way were they developed as prototypes of an economically viable commercial shipboard terminal.

A number of experimental and operational terminals have been developed for military ships. In the mid-1960's the U.S. Navy developed a series

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(10) Hanas, O.J. et al, "L-Band ATS-5-ORION-S.S. Manhattan Marine Navigation and Communication Experiment." Prepared for NASA, Electronics Research Center, Contract NAS 12-2260, by Applied Information Industries, Moorestown, New Jersey. June, 1970.

of 1.83 meter (6 ft.) diameter X-band shipboard satcom terminals known as the SSC-3's in conjunction with DOD's Initial Defense Communication Satellite Program (IDCSP). These employed precision 3-axis pedestals and were generally very complex systems. These terminals are currently (1971) being retrofitted for use with DOD's later programs. In addition, the U.S. Navy in 1970 commenced development of new X-band shipboard terminal sets which will be fully compatible with the Defense Satellite Communication System (DSCS).[11] Depending on the user ship class, these new sets (AN/WSC-2) will employ either 1.22 meter (4 ft.) of 2.44 meter (8 ft.) diameter parabolic antennas and will be even more complex than the SSC-3's. Because mast-mounting was found to create various interference and maintenance problems, each ship will employ two antennas, each side-mounted, with automatic switch-over between coverage sectors.

A relatively economical X-band military shipboard terminal system has been developed by Great Britain in conjunction with their SKYNET defense system. Referred to as the SCOT terminals, they employ an antenna diameter of 1.22 meters (4 ft.) and, with electronics for a single RF transmit and receive channel, are reported to cost about $240,000 per set.

In addition to the more complex, higher frequency communications electronics, such military systems are very costly due to the narrow beamwidths of even these small antennas at X-band (1-2 degrees) as well as a host of other unique features, such as automatic switch-over and operability while nearby large guns are being fired.

A closer parallel exists with the U.S. Navy's recently developed TacSatCom

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(11) Naval Electronic Systems Command - Contract Specification ELEX-S-41 Shipboard Satellite Communication Sets AN/WSC-2(XN-1)(V). 19 December 1969.

shipboard terminals which operate basically at high VHF frequencies. They are intended for narrowband voice and data applications. Two sizes were developed, one providing 7 dB of antenna gain and the other 12 dB.[12] While these gains and their corresponding beamwidths (approximately 78 degrees and 42 degrees, respectively) are representative of the civilian MARSAT* requirement, these antennas at the TacSatCom frequency are relatively large, requiring almost 30 times the aperture area than an antenna of equivalent gain at L-band. Both sizes are "single beam" antennas which are mounted on identical elevation over azimuth pedestals. In each case, antenna pointing is provided passively by slaving in azimuth to the ships gyrocompass, while manually locking the elevation axis in one of three (switched) angular positions. Realignment adjustments and calibration are typically required on a daily basis. The pedestal employed was developed to drive the relatively large antennas against high wind torques and yet still be very lightweight, and to meet the stringent Mil-E-16400 specification. The system was configured for operation with two antennas on-board each ship, with automatic switch-over functions incorporated in the servo system. They are thus quite costly. These Navy TacSat terminals are therefore not considered representative of devices attractive for the MARSAT application, although the pointing concept itself has features of definite interest.

The Navy is also currently in the process of initiating procurement of Fleet Broadcast terminals for another narrowband satellite relay system which operates near the upper end of the VHF band. These terminals will be used primarily for data communications and are to employ essentially omnidirectional (hemispherical coverage) antennas. Commercial

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* The acronym "MARSAT" is used interchangeably with Maritime Satellite for brevity throughout this report.

(12) Shipboard Antenna Systems for Use with the Tri-Service Tactical Satellite Communication System," Collins Type 837V-1 and 837W-1 Product Description, Collins Radio Company, Dallas (undated).

equivalents of such terminals could be expected to be quite attractive for shipboard use. However, as noted in Paragraph 1.1, the practical system economic trade-offs associated with the MARSAT application at L-band dictate the necessity of providing some higher value of antenna gain at the shipboard terminal.

Of course, NASA has developed shipboard antennas systems as well, but these antennas were designed for significantly more capability than is considered here.[13] Range ships have been used as part of the NASA's ground support networks as mobile telemetry stations, etc. Also special ships were instrumented for Apollo support, with large antennas and a full complement of S-band and VHF electronics. The NASA range ships are very specialized systems and the antennas used on them also are not representative of commercially practical solutions.

Apart from shipboard terminals, a number of analogous studies of L-band satcom antennas for aircraft have been performed in recent years, reflecting the great deal of attention given to the aeronautical satellite application. While the RF frequency bands used and gain requirements are approximately the same, the aircraft terminal analogy is really quite small. The severe dimensional size limitations and aerodynamic drag considerations greatly decrease the options available for aircraft terminals. Virtually all solutions proposed for aircraft involve some form of antenna array. Considerable effort has been expended in France on the development of L-band aircraft antennas in conjunction with Project Dioscures.[14] The Dioscures antennas developed by CNES and SGAC are a compromise between a horizontal fan beam and a pencil beam

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(13) NASA SP-87, "Proc. of the Apollo Unified S-Band Technical Conference," Goddard Space Flight Center, July 14-15, 1965.

(14) Manuali, B. "The Dioscures Project - System of Telecommunications, Air Traffic Control and Navigation by Satellites", Telecommunication Journal, Volume 36, February 1969.

array, in that switchable, non-symmetrical beams are formed by two flushed-mounted arrays on each side of the aircraft, providing a gain in the region of 9-12 dB. The method of controlling the beam-switching during aircraft motion in the Dioscures system is not known, although flight tests have been performed using computer-programed switching over prescribed tracks.

Further, NASA's Electronic Research Center sponsored a study of L-band steerable arrays for aircraft by TRW in 1968, which indicated the general complexities associated with achieving near-hemispherical coverage using phased arrays.(15) The study did not address the controlling mechanism for beam pointing. As will be discussed in later sections of this report, multiple arrays may be required to provide adequate constancy of gain over the greater-than-hemispherical coverage required.

Extensive studies of arrays have also been performed by Texas Instruments in conjunction with military aircraft systems.(16) These efforts have included development of prototypes, and TI has proposed a completely integrated system for commercial aeronautical satcom use. It incorporates an eight element linear array on each side of the aircraft, and two special (azimuth) hole-filler antennas for fore and aft signal directions. At the receive frequency of 1.55 GHz, the system has a maximum gain of 13 dB and a minimum gain of 9 dB. The array circuitry includes the formation of difference patterns, enabling beam pointing to be provided for by conventional monopulse tracking of the satellite signal.

While the environmental and dimensional constraints leading to the design of such arrays bear little relationship to the shipboard case, arrays in

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(15) Hering, Karl. "Parametric Study of L-Band Steerable Arrays for Aircraft." NASA-ERC Contract NAS 12-539, TRW Report No. 08710-6022-R0000. TRW Systems, Redondo Beach, Calif. September 1968.

(16) Roberts, Sam B. "Airborne L-Band Receive Array (U)", Technical Report AFAL-TR-70-136, U.S. Air Force Avionics Laboratory, Air Force Systems Command, Wright-Patterson AFB, Ohio. June 1970.

general have certain meritorious features, such as freedom from mechanical drives and/or amenability for fan-beam generations and will be considered in the subject maritime context in Sections 5 and 6.

The point of the preceding capsule survey of related efforts is that the user antenna subsystem for a commercial maritime satellite service is a fairly unique application which thus warrants a certain amount of basic conceptual investigation prior to drawing any specific conclusions in system planning or consideration of prototype development. This study therefore addresses the basic conceptual trade-offs associated with the design of the user antenna subsystem which, in volume production, would represent costs well over an order of magnitude less than the previously developed shipboard antennas noted above. A recurring cost in the range of $2,000 to $5,000 is subjectively considered to represent a viable value for a MARSAT user antenna subsystem, excluding installation and radio electronics.

1.3.3 Study Report Order

Consistent with the study objectives noted, the study report is structured as follows. Section 2 reviews the basic system performance factors and design constraints, noting the significance of shipboard antenna gain and beamwidth, defines alternative configuration categories and related performance criteria and reviews the various shipboard environmental constraints, such as multipath and vessel roll and pitch, etc.

Prior to discussing the various antenna radiator alternatives in Sections 4, 5 and 6. Section 3 reviews, analyzes and compares various techniques for beam pointing; including manual pointing, slaving to shipboard inertial references and conventional automatic tracking of satellite signals. Section 4 discusses "single-beam" antennas which are generally intended for mechanical pointing; Section 5 addresses multiple beam antenna arrays with beam switching and Section 6 considers electronically-steered array antennas. Section 7 then presents a summary of the analysis and a comparison of selected approaches, and Section 8 contains pertinent conclusions and recommendations.

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SECTION 2

PERFORMANCE FACTORS AND DESIGN CONSTRAINTS

2.1 SIGNIFICANCE OF SHIPBOARD ANTENNA GAIN AND BEAMWIDTH

Prior to the consideration of the complexities of various shipboard antenna subsystems, it is of value to review briefly the basic significance of user antenna gain and related implications of antenna beamwidth. The most fundamental system design trade-off in configuring the maritime satellite service is that between three basic opposing goals:

1. Minimization of required total satellite EIRP
2. Maximization of system service capacity, quality and availability
3. Minimization of user terminal complexity.

Central to all three of these goals is the fact that system cost effectiveness forces the system to be severely power-limited in the satellite-to-ship path. The basic satellite payload which is primary in establishing the cost of the system's space segment is that of satellite weight, which is directly related to satellite transmitter power.

2.1.1 Satellite Power Utilization and System Capacity

The satellite-to-land based (coast) station link can be established cost-effectively with large coast terminal antennas, since these stations would be few in number, and it is the transmitter used for the satellite-to-ship link which requires the larger percentage of available dc power produced by the satellite's solar cell arrays. (The other two links of the four-link system, those to the satellite from user ships and from coast stations, obviously require negligible satellite dc power.) The notion of two separate satellite transmitters is predicated on the requirement that the MARSAT system be configured for full duplex

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operation, which is assumed for purposes of this study. Due to the subject power limitations, spectrum constraints and related factors, this assumption may be extended to the conclusion that the satellite transponder should be cross-strapped with another RF band (e.g., SHF) for use in the satellite-to-coast and coast-to-satellite links, so that reference herein to the satellite's L-band transmitter signifies that for the satellite-to-ship link only.

Thus, with the L-band transmitter sizing directly affecting the space segment cost, system cost-effectiveness demands that the link be sized to provide a certain capacity and desired quality most of the time, but not necessarily all of the time. Even though recent maritime mobile spectrum allocation at L-band is probably the best space link frequency in the entire spectrum from dc to light in terms of freedom from propagation anomalies, there are sources of dynamic range in received signal levels, including a small amount of fading due to atmospheric effects and multipath. It would thus be wasteful to compromise design capacity to insure that the system provided desirable quality during all adverse conditions; rather, it is necessary to permit a degradation to an acceptable quality during adversities which exist a small percentage of the time.

Of the basic satellite services required - radiotelephony (voice), data (telex, facsimile) and eventually position fixing (navigation), it has been shown that it is the voice services which require the larger portion of satellite transmitter power.[3] It has accordingly been derived that a credible service specification is constituted by the requirement that the system be sized to provide an articulation index (AI) of about 0.50 to 0.55 (i.e., 0.53), 95% of the time, while the ship-to-satellite elevation angle is 15 degrees.[4] This value of AI (an empirical parameter relating to signal processing factors and received signal-to-noise ratio)

(3) 1970 DOT-CG Study, Volume III. Op. Cit, pg. 6.

(4) 1971 DOT-TSC Study, Op. Cit. pg. 14.

can be loosely considered to correspond to a word intelligibility of some 93-94% (256 word list) or a sentence intelligibility of some 97-98%.[17] Given certain assumptions on L-band propagation fading characteristics and modem performance, the above service specifications can be shown to also provide for an AI of 0.3 or better, 99.99% of the time (0.3 AI is frequently considered as a threshold in intelligibility). It can also be shown that for certain optimum voice modems (e.g., phase-locked NBFM and coded four-phase DM/PSK), the above service specification corresponds to a received carrier-to-noise density (C/N_0) at the shipboard terminal (IF) of about 45 dB-Hz, 95% of the time. (Figure 1-6)

Given this data point, along with various other assumptions, the basic trade-off between the three basic objectives listed above can be presented as in Figure 2-1, which shows total satellite transmitter power required as a function of shipboard terminal G/T and satellite capacity, i.e., number of channels, where each channel is of voice capacity - i.e., producing a C/N_0 of 45 dB-Hz, under the stated conditions. The terminal G/T is the ratio of antenna gain G to the system noise temperature Ts (in degrees Kelvin) at the receive frequency of 1540 MHz and where both parameters are referenced to the same point in the receive path (typically the input connector to the preamplifier). The relationship between shipboard antenna gain and G/T is indicated in Figure 2-2 for two baseline values of system noise temperature (at an elevation angle of 15 degrees), depending on which class of low-noise preamplifier is used.

2.1.2 Baseline Satellite-to-Ship Path Link Parameters

It is not the intent to dwell on complete link budget details in this report, but obviously Figure 2-1 is based on certain assumptions. The baseline power budget for the satellite-to-ship link used as a reference is shown in Table 2-1. Most of the items are self-explanatory; however, a few comments are necessary. For purposes of this report, the budget is

(17) Philco-Ford TR-DA 1583 (III), "Systems Engineering Study of Aeronautical Satellite Services - Final Report to COMSAT Corporation", Philco-Ford Corporation, Palo Alto, California. 15 December 1967.

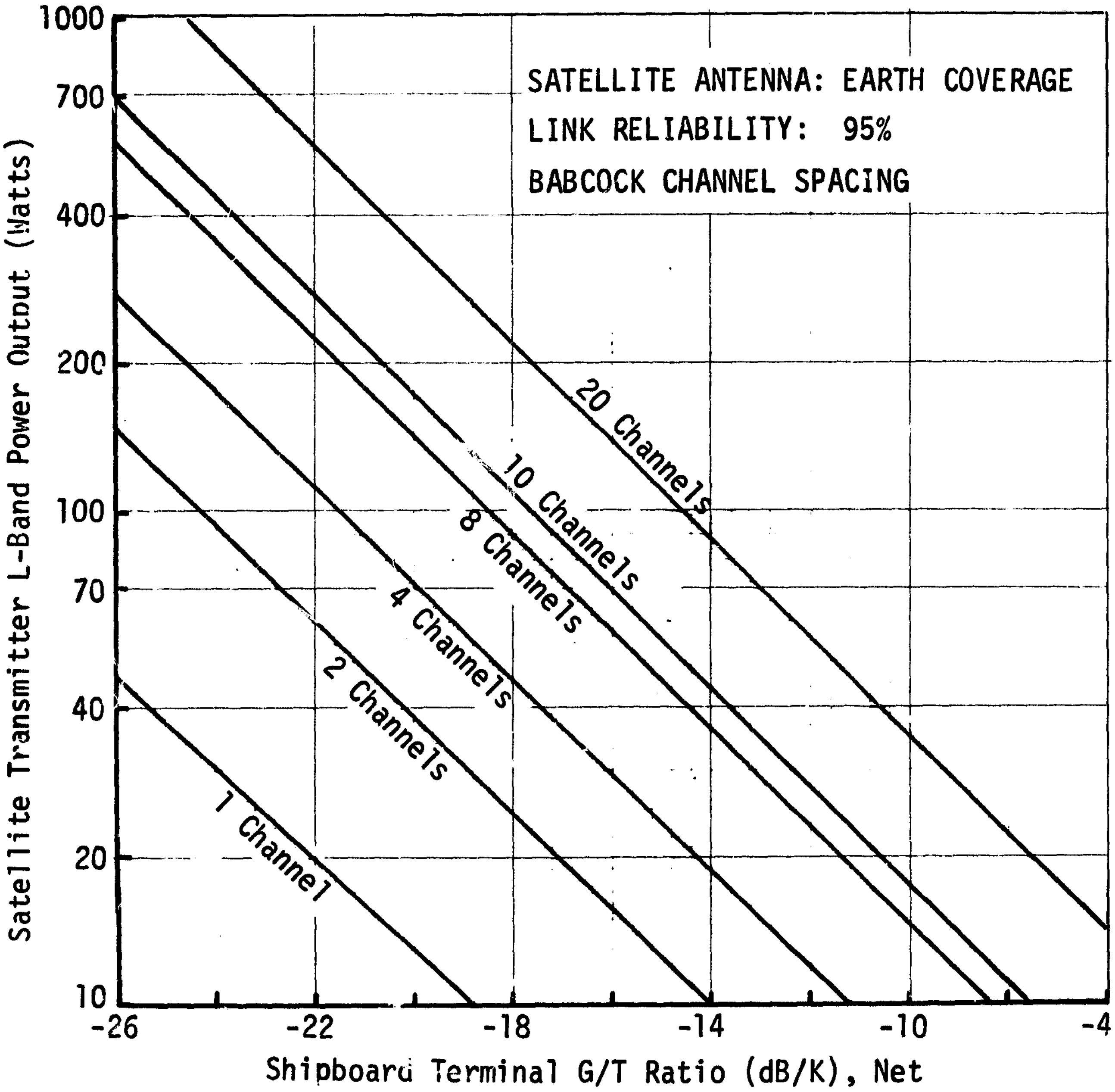

Figure 2-1 SATELLITE RF POWER VS USER TERMINAL G/T AND SYSTEM CHANNEL CAPACITY

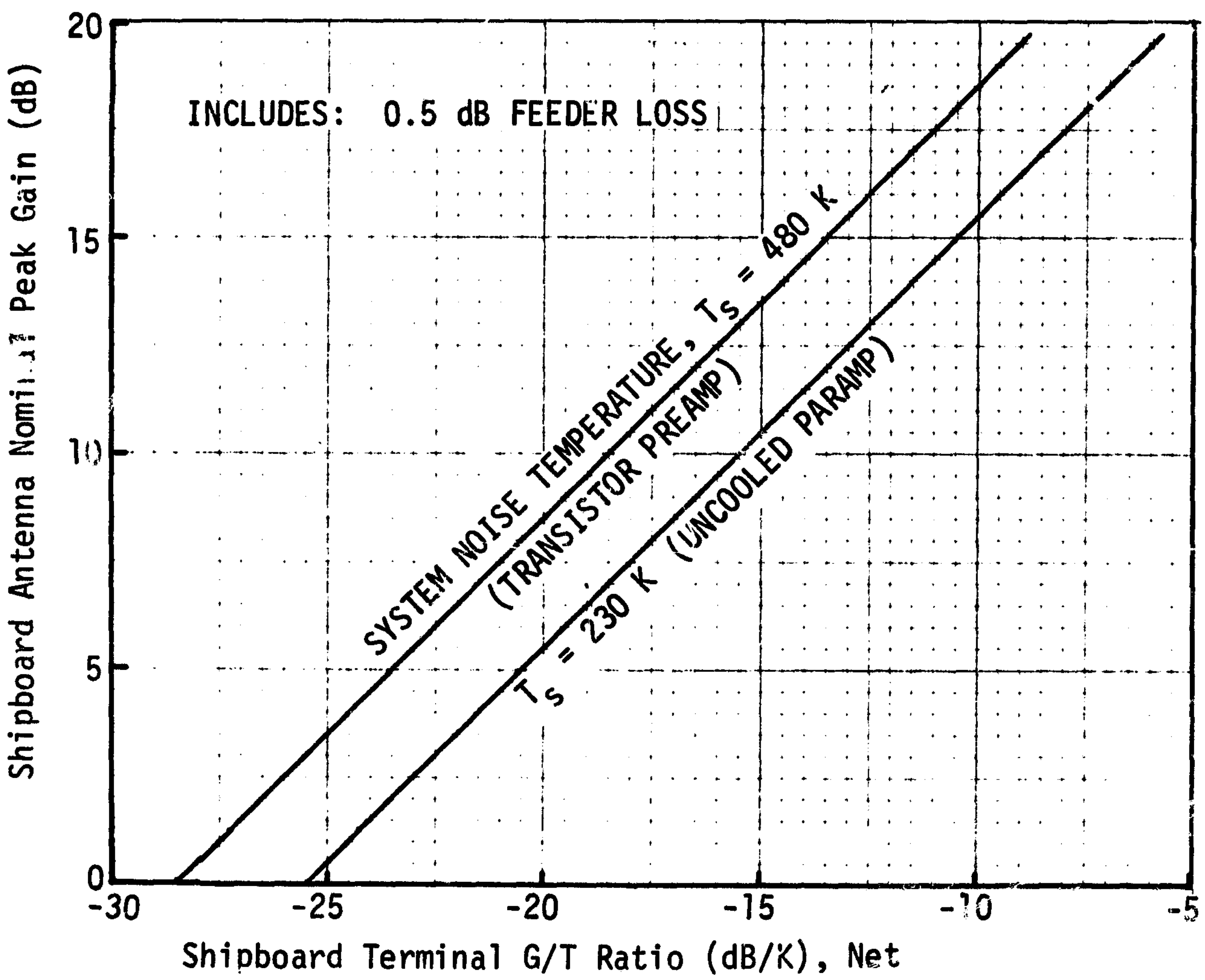

Figure 2-2 SHIPBOARD ANTENNA NOMINAL GAIN VS G/T FOR SPECIFIC RECEIVE SYSTEM NOISE TEMPERATURES

TABLE 2-1 BASELINE SATELLITE-TO-SHIP LINK BUDGET AT 1540 MHz

Item	Link Parameter	Allocation
1.	Satellite Transmitter Output Power per Channel	+ 9.8 dBW
2.	Satellite Antenna Coupling and Line Loss	- 0.5 dB
3.	Satellite Antenna Gain - Peak	+18.6 dB
4.	Satellite Antenna Gain Off Beam-Center Loss	- 2.5 dB
5.	Satellite EIRP	+25.4 dBW
6.	Free Space Attenuation	-188.3 dB
7.	Propagation Loss (Scintillation, Absorption, Multipath)* 95% Reliability	- 1.2 dB
8.	Polarization Loss*	- 0.8 dB
9.	Shipboard Antenna Gain, Peak	+10.0 dB
10.	Ship Antenna Gain Off-Center Loss	- 1.3 dB
11.	Shipboard Receive Line Loss	- 0.5 dB
12.	Received RF Signal Level	-156.7 dBW
13.	Shipboard Receiver System Noise Temperature*	26.8 dB-^{o}K
14.	Boltzmann Constant	-228.6 dB/^{o}K-Hz
15.	Receive System Noise Power Density	-201.8 dBW/Hz
16.	Up-Link Noise Degradation	- 0.1 dB
17.	Received C/N_o per Channel (Voice Capacity)	45.0 dB-Hz

* Referenced to 15^{o} user-to-satellite elevation angle

based on the specified service requirement of a received C/N_0 at the user terminal of 45 dB-Hz per channel and an arbitrary assumption of a nominal antenna gain of 10 dB at the receive frequency. The budget is then "solved" for satellite transmitter output power required per channel, which is seen to be about 9.5 watts (+9.8 dBW). It assumes a single satellite L-band antenna with an earth coverage pattern. Considering the distribution of ships at sea, total earth coverage is a firm necessity for a maritime services satellite (in contrast to the aviation case where the majority of traffic is contained in smaller regional sectors). Higher satellite antenna gain is theoretically possible with steerable or selectable narrow-beams, but consideration of such spacecraft complexity is not considered appropriate in this study.

The propagation loss assumed for a 95% probability of not being exceeded is 1.2 dB at an elevation angle of 15 degrees. This is consistent with recent opinions at the ICAO-ASTRA IV meeting in Montreal.(18) Twice this value, i.e., 2.4 dB, might be appropriate for a 99% probability point. Figure 2-3 shows the effect of specified link reliability on required user terminal G/T, assuming a fixed capacity of eight channels. L-band propagation phenomena are still quite controversial, and it may be possible that larger values of fading could occur at certain times in certain small equatorial regions.(19) However, in most regions, most of the time, fades would be expected to be less than 0.3-0.5 dB.(20)

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(18) ASTRA IV Report to the Air Navigation Commission of the International Civil Aviation Organization (ICAO) on a meeting January 11 to 22, 1971. Montreal, Canada.

(19) Crampton, E.E. and Sessions, W.B. "Experimental Results of Simultaneous Measurement of Ionospheric Amplitude Variations of 136 and 1550 MHz Signals at the Geomagnetic Equator, Westinghouse Electric Corporation, Report 7, January 1971.

(20) Kissel, F.J., "Earth/Space Propagation at L-Band Using the ATS-5 Spacecraft," Proc. of the IEEE International Conference on Communications, San Francisco, June 9, 1970.

The primary source of the above fades would probably be multipath. Considering that this effect is less for ships than for aircraft, the assumed value is considered sufficiently conservative.

The assumption of a shipboard antenna pointing loss of 1.3 dB is quite arbitrary. It might be considered as representative of a compromise between a steerable, tracking antenna and a switched-beam fixed array. In this context, it is important to note that the values of terminal G/T in Figure 2-1 are net values in the direction of the satellite, i.e., operational values. For example, the operational value of antenna gain in Table 2-1 is 8.7 dB, and referred to the preamplifier input, it is only 8.2 dB, even though the antenna radiator itself may have a peak gain of 10 dB. With a system noise temperature of 26.8 dB-K, the operational G/T of Table 2-1 is then 8.2-26.8, or -18.6 dB/K. Thus, the G/T values used in Figures 2-1 and 2-3 are operating net values which include the effect of pointing loss as well as line loss.

The shipboard receiver system noise temperature is based on an antenna temperature of 100 K and on the use of a low noise transistor preamplifier with a noise figure of 3.5 dB, which is considered practical for the near future. The uncooled parametric amplifier curve in Figure 2-2 is for a noise figure of 1.3 dB.(3) The final factor, which is not listed in the table, is the multiple-channel sharing factor. Table 2-1 is for a single channel with capacity for voice or data at about 1200 bps-2400 bps (depending on bit error rate allowed). Sizing the system for multiple (N) channel capacity - such as eight channels does not simply increase the transmitter power required by N, e.g., 8 (9 dB). There is a certain amount of loss associated with the multiple access of a single power-limited transmitter power amplifier in the satellite. The amount depends on the accessing scheme, whether digital (TDMA) or FDMA or FDM/PM, etc., and the number of channels; and, in the case of FDMA,

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(3) 1971 DOT-TSC Study, Op. Cit. Section 15.

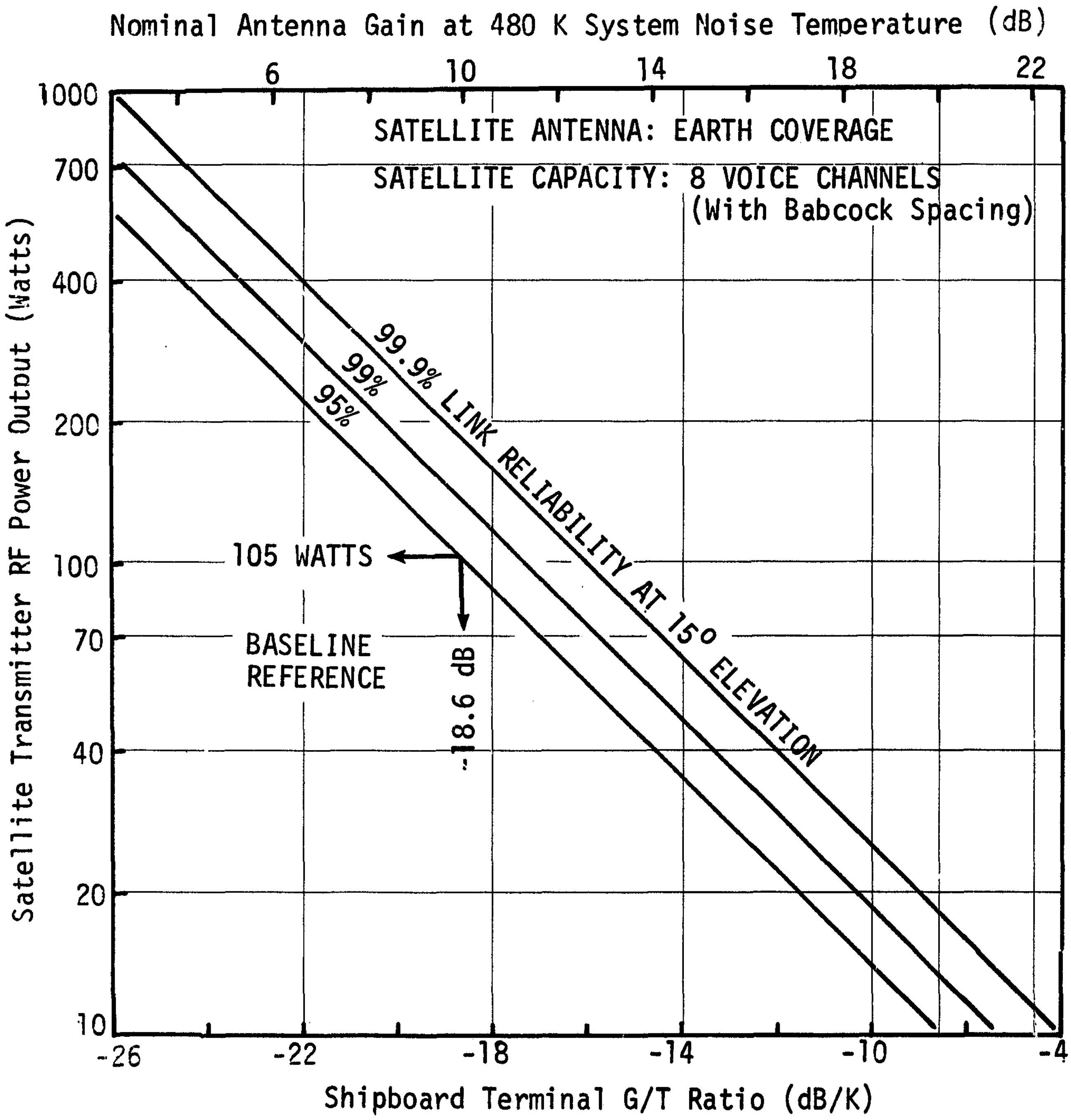

Figure 2-3 SATELLITE RF POWER VS USER TERMINAL G/T AND LINK RELIABILITY

also on the channel (carrier) spacing. While time division multiplexing of digital signals (including voice) has the least multiple access loss, it is relatively inflexible in terms of independent coast terminal use and requires digital voice modems which may or may not be as efficient for individual channels. Table 2-2 shows the trade-offs involved between FDM access schemes.(3) For purposes of Figures 2-1 and 2-3, conventional FDMA was assumed with Babcock intermodulation-controlled channel spacing, which yields high power efficiency, access flexibility and user terminal simplicity at the expense of spectrum efficiency. This is considered most appropriate for an initial MARSAT system, as noted in Paragraph 1.2.1, permitting easy growth into a much higher capacity second generation system.

It should be clear from Figures 2-1 and 2-2 that a small amount of shipboard gain is necessary to provide some system capacity with realistic satellite L-band transmitter power amplifiers. If an eight channel capacity is considered as a bare minimum for a viable initial system,(4) it is seen that a nominal user antenna gain of about 10 dB requires a satellite transmitter power amplifier designed for about 105 watts peak output. This, in turn, can be translated into spacecraft end-of-life total dc power of about 435 watts, which approaches a limit for a Thor-Delta launched spin-stabilized satellite, assuming certain extrapolatable capabilities for the Thor-Delta class in the 1975 time frame. A three-axis stabilized satellite could more than double this value, which would relax the user antenna gain requirement by almost 4 dB (e.g., from 10 dB to 6 dB) for the same capacity requirement (i.e., 8 channels).(4)

2.1.3 Example Antennas and Relative Beamwidth

It may also be seen that a relatively small change of user gain could significantly alter the revenue producing capacity of the satellite.

(3) 1970 DOT-CG Study, Op. Cit.
(4) 1971 DOT-TSC Study, Op. Cit., Volume III.

TABLE 2-2 RELATIVE POWER AND RF SPECTRUM EFFICIENCY OF MULTIPLE AND SINGLE CARRIER FDM ACCESS SCHEMES

NO. OF CHANNELS n	FDMA BABCOCK SPACING		FDMA ADJACENT SPACING		FDM/PM SINGLE ACCESS	
	La, dB*	$\frac{RF\ BW}{Bch}$	La, dB	$\frac{RF\ BW}{Bch}$	La, dB	$\frac{RF\ BW}{Bch}$
4	1.46	7	2.6	4	2.7	8
6	1.36	18	3.3	6	2.7	12
8	1.26	35	3.9	8	2.7	16
10	1.25	62	4.2	10	2.7	20
20	1.25	$\sim 10^3$	4.4	20	2.7	40

* La is the net system loss due to multiple access, relative to the ideal accessing of n channels.

This may seem relatively insignificant since a number of antenna radiators are feasible in this range which are all fairly small. Especially in the range 7-14 dB at 1540 MHz, the antennas can be quite small relative to ship geometry. A simple cavity-backed spiral will give a nominal gain of 7 dBi (dB above isotropic) with good circular polarization and is only about 15 cm (3.3 inches) in depth, and weighs less than 0.9 kg (2 lbs). Clearly, this is a very small antenna, and since cavity-backed spirals are readily amenable to fiberglass covering, such a selection would have many attractive features. Similar performance could be achieved with a small helix about 20 cm (8 inches) long and 11 cm (4.5 inches) in diameter. Larger helices can be used to provide higher gains, e.g., 66 cm (26 inches) long by 23 cm (9 inches) in diameter can provide a peak gain of 15 dBi at 1540 MHz.

Another particularly interesting radiator is the "short-backfire" antenna which has a high (70%) efficiency, producing over 14 dB of gain at 1545 MHz. This antenna is shaped like a pie plate, 39 cm (15.4 inches) in diameter, with a rim about 4.9 cm (1.9 inch) deep and a small (10 cm diameter) feed protruding out another 4.9 cm.

The three example radiators above can all be foam filled as well as radome covered with fibreglass or plastic and readily made corrosion resistant. Their characteristics are described along with those of other candidates in Section 4. The point is that with respect to antenna radiator size, there is little significance to the difference of a few dB of gain. With the gain being so valuable to satellite system performance, there would appear to be strong motivation to use the slightly larger antennas.

However, higher gain by definition means narrower beamwidth. Beamwidths less than that of a hemispherical pattern mean that multiple radiators must be used, or that a single radiator be directable in orientation in some manner. More crucially, a mechanism must be provided for in the shipboard terminal for <u>deciding</u> which beam position is best. This could be an automatic satellite angle tracking receiver, or a calibrated inertial reference, or an operator who monitors and controls the beam position. The most suitable scheme, and its complexity depends to a great degree on the actual value of beamwidth, i.e., gain.

It may readily be shown[21] that the beamwidth of an antenna is related to its gain G by the following:

$$\theta_h \theta_v = \frac{k_g}{|G|} \qquad (2\text{-}1)$$

where θ_h and θ_v are the angular beamwidths in the horizontal and vertical planes, respectively, (relative to some radiator geometric reference) and $|G|$ is the absolute (numerical power ratio) gain. For a conventional antenna with a symmetrical beam pattern,

$$\theta_h \theta_v = \theta^2_{hp} \qquad (2\text{-}2)$$

where θ_{hp} is the two-sided, half-power (3 dB) beamwidth of the main pattern. The parameter k_g in Equation (2-1) is considered as a constant in the region of 30,000 when Θ_h and Θ_v are in units of degrees. In fact, the value of k varies slightly as a function of various antenna design parameters, including efficiency and energy distribution. For a typical pencil-beam (high gain) antenna of average (50-60%) efficiency, the value of k_g is about 26,600. This relationship is fairly accurate for gains in excess of 10-15 dB. As antenna gain becomes lower than this, the value of k_g increases somewhat, e.g., to 30,000. Also, it varies somewhat between antenna types even for the same gain. Figure 2-4 demonstrates this approximation.

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(21) Silver, S. "Microwave Antenna Theory and Design." McGraw-Hill Book Company, New York, N.Y. 1949. (For example, $G = \eta\pi^2(d/\lambda)^2$ and $\Theta_{hp} = k_b(\lambda/d)$ rad, or $(360/2\pi)k_b(\lambda/d)$deg. where η = efficiency, d = diameter and k_b = a beam shape factor near unity; yielding $G = 32{,}500\ \eta k_b^2/\Theta_{hp}^2$)

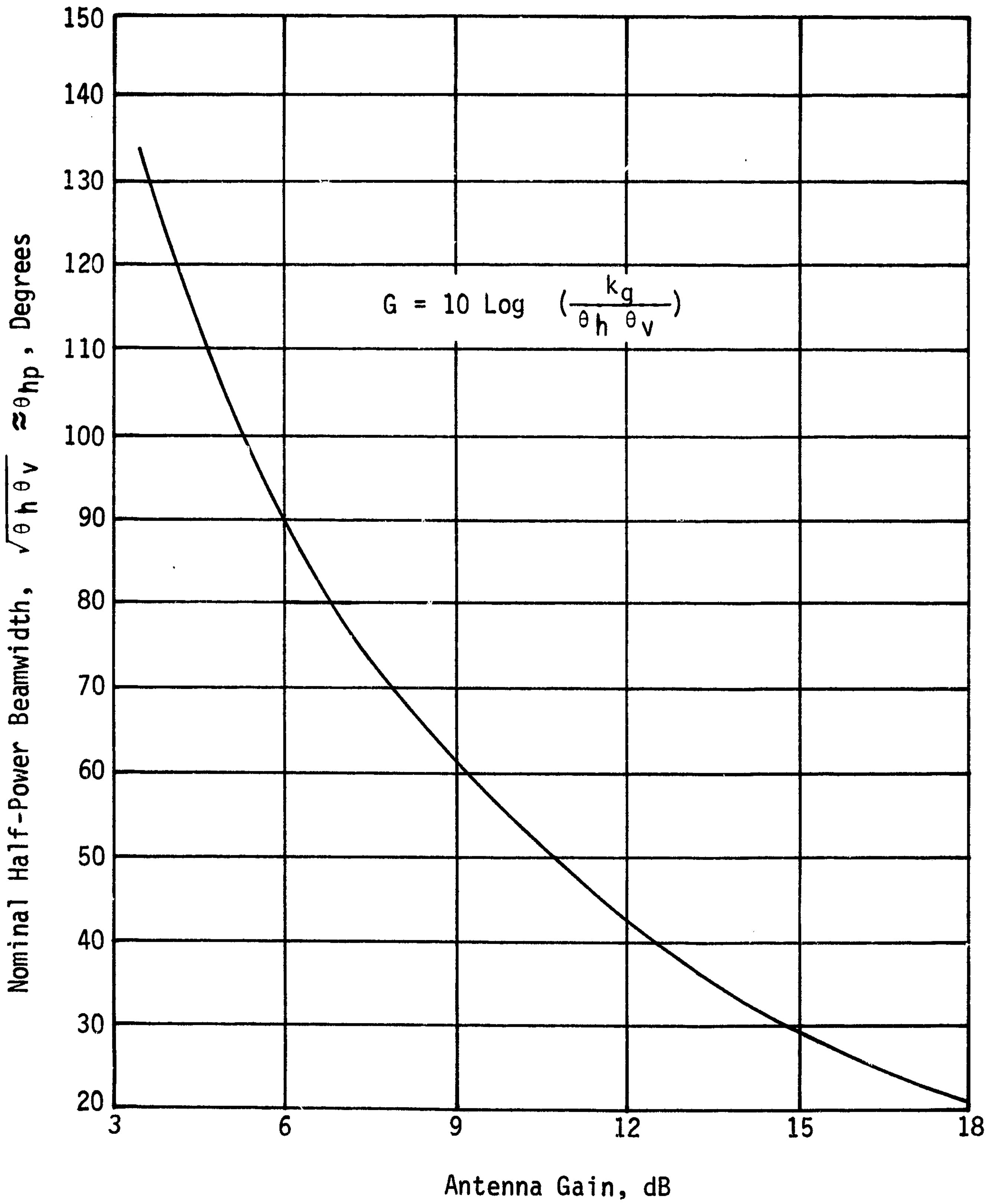

Figure 2-4 ANTENNA GAIN AND BEAMWIDTH

Thus, while increasing the shipboard user antenna gain would appear to be a trivial imposition in terms of radiator size, the decreased beamwidth clearly increases the complexity of the terminal system by imposing more stringent pointing requirements. An 18 dB gain antenna, with a half-power beamwidth of about 21 degrees, would need to be pointed to within ±10 degrees continuously to insure an operational antenna gain of 15 dB. Since the satellite is effectively stationary in the synchronous equatorial orbit, it is strictly the ships motion which must be compensated.

As is described in Section 2.4, even the very large ship can roll ±30 degrees in heavy seas. Small ships, like cutters, can roll ±50°. In addition to roll, angular dynamics are caused by pitch and yaw, and heave, surge and sway. Even the ships intended motion changes the apparent satellite direction, by as much as 8 degrees in one day at a ship speed of 20 knots.

With a very wide beam antenna, compensation for motion could be made very easy, of course. A perfect hemispherical pattern (while not trivial to design) could be used to provide a gain of 1-2 dB and negate entirely any pointing problem. Any more gain would produce some directivity. However, for low gains such as 3-6 dB, the beam would generally be wide enough to accommodate all dynamics due to heavy seas, although an operator would still have to manually orient the antenna towards the satellite on a periodic basis, e.g., every few days. As gain becomes larger, manual tracking becomes less and less practical, since it would not be a feasible scheme to compensate for heavy sea motion.

Thus, for even the moderate gains of 8-10 dB or more, some form of continuous beam control must be mechanized for the ship terminal. Considering a shipboard subscriber population on the order of 10^4, if the required pointing mechanism costs much over $10,000 per unit, then it probably isn't cost effective in that the additional >$100 M might be better spent in the space segment. However, if it costs less than a few thousand dollars, it would be a cost-effective requirement for the user terminal.

2.1.4 Antenna Gain and Shipboard Transmitter Power

There is one aspect of shipboard gain which is not directly affected by satellite system capacity. Independent of the number of channels used for the system, each ship will transmit only one RF carrier at a time. Of the potential voice, data and navigation services to be available, it will be the voice service which requires the most power, even in the up-link (from the ship to the satellite). The satellite's receiver system noise temperature would not in practice be improvable by using a larger satellite. Given a fixed gain limitation on the spacecraft antenna, such as earth coverage, the up-link power level required is effectively a constant, independent of satellite complexity or capacity. Thus, the transmitted EIRP required of the shipboard user is effectively a constant. Table 2-3 contains a baseline link budget for the ship-to-satellite path, where is may be noted that a user EIRP of +28.3 dBW net (almost 700 watts) is required in the direction of the satellite, in order to insure a modest carrier-to-noise density ratio in the satellite transponder. While the link from the satellite-to-the coast station can be configured to add little noise to the ship transmissions, a higher quality channel (in terms of audio SNR; not necessarily intelligibility) is required in this direction to interface properly with domestic carriers.[4]

Thus, whatever the shipboard user antenna gain, a user EIRP of about 1 kW is required. (Since this link is not so critically power-limited, it is considered appropriate to use a 99% probability of achieving the desired high quality at an elevation angle of 15^{o}.) With a high gain antenna, such as 16 dB, the transmitter power amplifier device could be a simple transistor device rated at 25 watts. This could easily be mounted at the antenna for low transmit line loss. As noted in the

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(4) 1971 DOT-TSC Study, Op. Cit.

TABLE 2-3 BASELINE SHIP-TO-SATELLITE LINK BUDGET AT 1640 MHz

Item	Link Parameter	Allocation
1	Ship Transmitter Power per Channel (100 watts)	+20.0 dBW
2.	Ship Antenna Coupling Loss	- 0.8 dB
3.	Ship Antenna Gain - Peak	+10.4 dB
4.	Ship Antenna Off-Beam Center Loss	- 1.3 dB
5.	Shipboard EIRP	+28.3 dBW
6.	Free Space Attenuation*	-188.7 dB
7.	Up-Link Propagation Margin for 99% Reliability*	- 2.4 dB
8.	Polarization Loss	- 0.8 dB
9.	Satellite Antenna Gain - Peak (19° Beam)	+19.0 dB
10.	Satellite Coverage Off-Beam Center Loss*	- 2.5 dB
11.	Satellite RF Coupling Loss	- 1.1 dB
12.	Received RF Signal Level	-148.2 dBW
13.	Satellite Receiving System Noise Factor a) Receiver Noise (transistor) $360^{\circ}K$ b) Antenna and Line Noise (Satellite over cold ocean) $190^{\circ}K$ $\}$ $550^{\circ}K$	-27.4 dB-$^{\circ}K$
14.	Boltzmann Constant	+228.6 dBW/$^{\circ}K$-Hz
15.	Receive System Noise Power Density	-201.2 dBW/Hz
16.	Nominal C/N_0 into Satellite Transponder	+ 53.0 dB-Hz

* Referenced to 15° user-to-satellite elevation angle

baseline budget, use of a nominal 10 dB (receive) gain antenna dictates the need for a 100 watt transmitter power amplifier. This could also be achieved in solid state, using a series/parallel network of L-band power transistors. However, use of a very low gain antenna (e.g., 3-6 dB) would require a transmitter power amplifier rated at 500-1000 watts. Figure 2-5 shows the shipboard transmitter power required as a function of antenna (receive) gain used. (The case for a satellite receiver using a low noise parametric amplifier is included for reference, even though such use is not current).

Imposing a requirement for transmitter power amplifier rating much higher than 100 watts would necessitate the use of an L-band travelling-wave tube (TWT). While the costs of such transmitter tubes can be expected to decrease over the next few years, especially in volume use, it would also be expected that a 500-1000 watt TWT, with its attendant support and cooling equipment might cost $2000 to $5000 more than a 100 watt solid-state device. In addition, its complexity would negate the practicality of its being located at the antenna, so that a high L-band transmission line loss would have to be compensated for by even higher power amplifier ratings.

Thus, there is even stronger motivation for the use of a moderate amount of antenna gain in the shipboard terminal. The above rationale indicates the potential cost effectiveness of the antenna beam pointing mechanism, even if it costs somewhat more than the few thousand dollars originally considered. Obviously, of course, the pointing scheme should be selected for as small a cost as possible, but it is noteworthy that its value to the Maritime Satellite system concept can be more significant than it appears on the basis of satellite power utilization only.

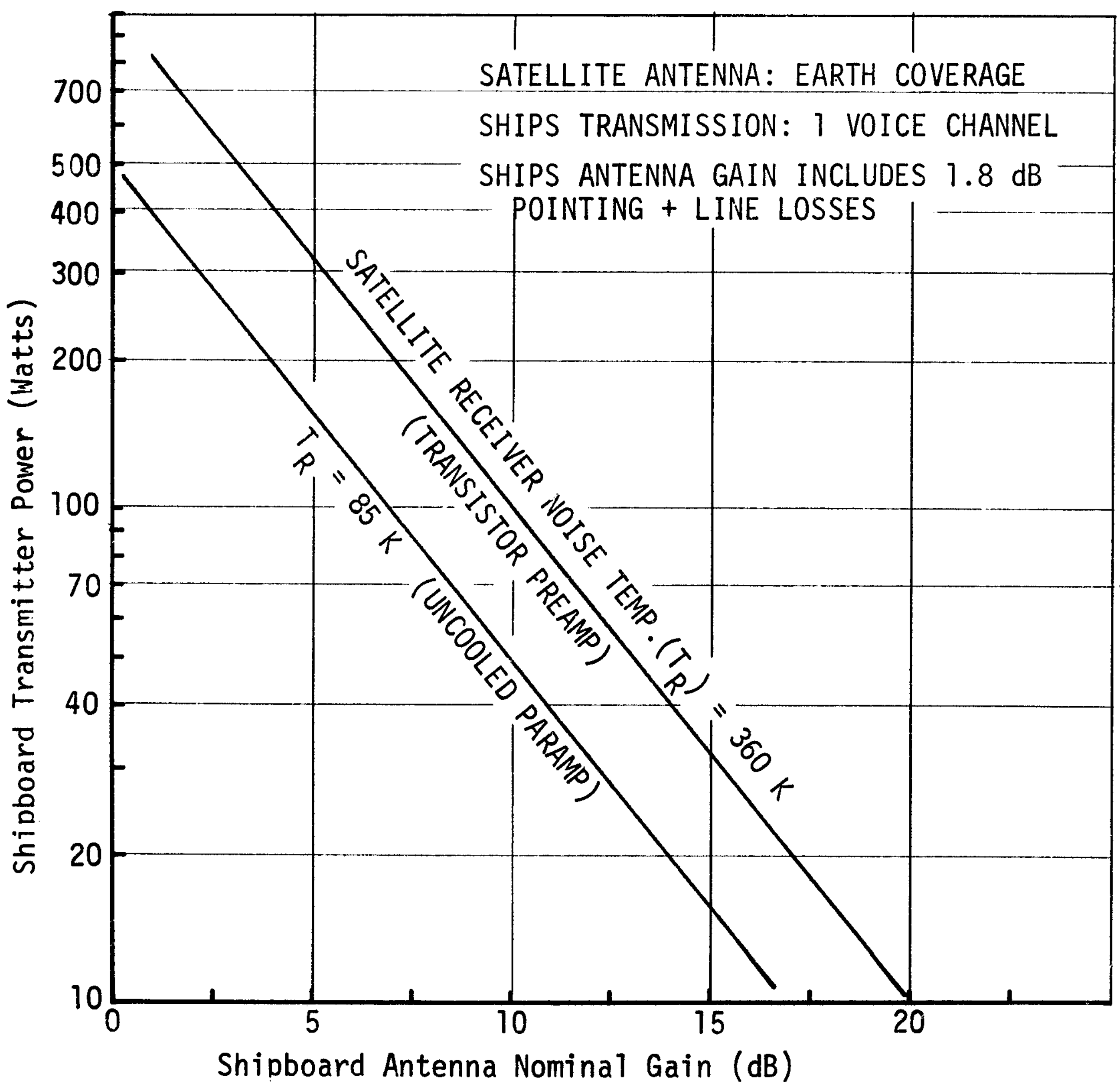

Figure 2-5 SHIPBOARD TRANSMITTER POWER VS NOMINAL ANTENNA GAIN FOR SPECIFIC SPACECRAFT RECEIVER NOISE TEMPERATURES

2.2 GENERAL ANTENNA SUBSYSTEM CONFIGURATION ALTERNATIVES AND PERFORMANCE FACTORS

2.2.1 Configuration Categories

1600 MHz is a relatively convenient band for antenna design, in that there are many options available and mechanical tolerances are not difficult, especially in the range of gain 3 dB to 18 dB generally considered for the MARSAT user application. A wide range of radiators are practical, each of which has certain attractive features. Most radiators may be used in arrays, where beam selection can be accomplished by simple switching. Single beam forming arrays can be used, where the beam is electronically steered using controlled phase shifters.

Rather than separate the various antenna possibilities by gain or by radiator design, it is appropriate for the purposes of this study to categorize antenna configuration alternatives by their implied beam-pointing scheme. Accordingly, the following three categories are defined:

1. "Single-beam" antennas which are mechanically pointed.

   This category is directed at conventional single element radiators which are driven by remote control of motors. However, it also includes array antennas which form a single beam, such as a linear or planar array of crossed-dipoles, for example, where the whole array is mechanically driven. In either case, the mechanical drive may be controlled by a human operator, or remotely by slaving to inertial references, or by an automatic angle-tracking receiver of various types. Generally, this category applies to use of a single mast-mounted antenna subsystem.

2. Multi-beam antennas with beam-switching.

This category is for antenna subsystem configurations which consist of an array of independent radiators arranged to cover the hemisphere by placing of beams adjacent to each other. The antennas may or may not be located next to each other, i.e., the category includes configurations such as a "ring" of adjacent radiators around a mast, or well-separated antennas strategically located around one of the upper decks. Beam selection is by switching, where the switching itself may be controlled manually, or by slaving, or by automatic pattern-null tracking, as in Category 1.

3. Electronically-steered beam antennas.

   This category is constrained to those antennas in which an array of elements is configured to produce a single beam which is electronically scanned by the continuous control of precision phase-shifters. As noted in Section 1, the phased-array category is that commonly considered for aircraft in the aeronautical satellite application. As in the other categories, the array phasing can be manual, slaved or autotrack-receiver controlled.

These categories are analyzed in Sections 4, 5 and 6 respectively, following an independent review of the pointing sensor mechanism in Section 3 and a discussion of general performance and design criteria and constraints in the remainder of Section 2.

2.2.2 Performance and Design Considerations

Antenna studies traditionally are directed virtually at a single parameter, aperture efficiency. For the MARSAT application, this parameter is at best secondary, and there are a wide variety of factors which are considerably more significant. An important consideration for example is location, in the sense that certain antennas preclude mast-mounting of critical components such as power amplifiers and low-noise preamplifiers.

Another is whether or not any inherent limitations exist in achieving a certain polarization. One potentially attractive antenna subsystem approach would be that of a fan-beam, which had an elevation beamwidth sufficient to accommodate all ship's roll angles, and then slaved in azimuth to the ships gyrocompass. Such an antenna could be quite simple; e.g., considerably less complex than the radar antenna mast-mounted on ships. However, the total elevation beamwidth required must include the total roll amplitude. For the very large ships, this might be 150°, (90°±30°), and for medium sized ships up to 180°. If a gain of 10 dB is desired, for example, the beamwidth product ($\Theta_h \Theta_v$) must be about 2800 as may be noted in Figure 2-4. This means the azimuth beamwidth must be on the order of 15° - 18°. From the standpoint of gyrocompass repeater accuracy and calibration frequency, this would appear credible. But beamwidth ratios of 10 are very difficult to achieve; for most radiator designs, such as horns, a practical limit is 4 or 5. A linear array might be used with more complexity, but in either case, achieving good circular polarization <u>and</u> a high beamwidth ratio is difficult. This fan-beam concept also suffers from the fact that the roll component of ships motion does produce a significant azimuth angle error when the satellite is in the direction of the bow-stern line, for which the gyrocompass would not compensate.

It is considered of value to delineate here some of the various factors and parameters which must generally be considered in the selection of antenna configurations. These are as follows:

1. Relative multipath effects
2. Polarization limitations

3. Swept volume and weight
4. Any practical gain or beamwidth limitations in the range 3 dB - 18 dB
5. Amenability to greater than hemispherical coverage by suitable steering
6. Dimensional size characteristics, location limitations
7. Amenability to locating low-noise preamplifier(s) and transmitter power amplifier(s) at feed port
8. Amenability to developing a pattern null to facilitate automatic tracking
9. Suitability for developing fan-beams and practical beamwidth ratio limitations
10. For phased arrays, such critical factors as element spacing, tilt angle, transmission line component losses, etc.
11. Any inherent requirement for multiple receivers and/or transmitters
12. Interference properties, sidelobe and backlobe levels
13. Relative noise temperature
14. Relative cost of development and manufacture
15. Suitability to radome protection
16. Rotary joint and/or cable-wrap implications
17. Wind-loading and vibration effects
18. General maintainability
19. Crew safety
20. Life

These considerations are addressed where appropriate in the following sections with respect to individual design concepts. Two especially significant general design considerations are that of multipath and the severity of the physical environment of ships at sea. These are reviewed next.

2.3 MULTIPATH DEPENDENCE

The performance of any communications system can be affected by multipath fading to some extent, and precautions must be taken to minimize the effect.

Two 'rays' are received at the antenna at any time, a 'direct ray' and a 'reflected ray' as shown in Figure 2-6.

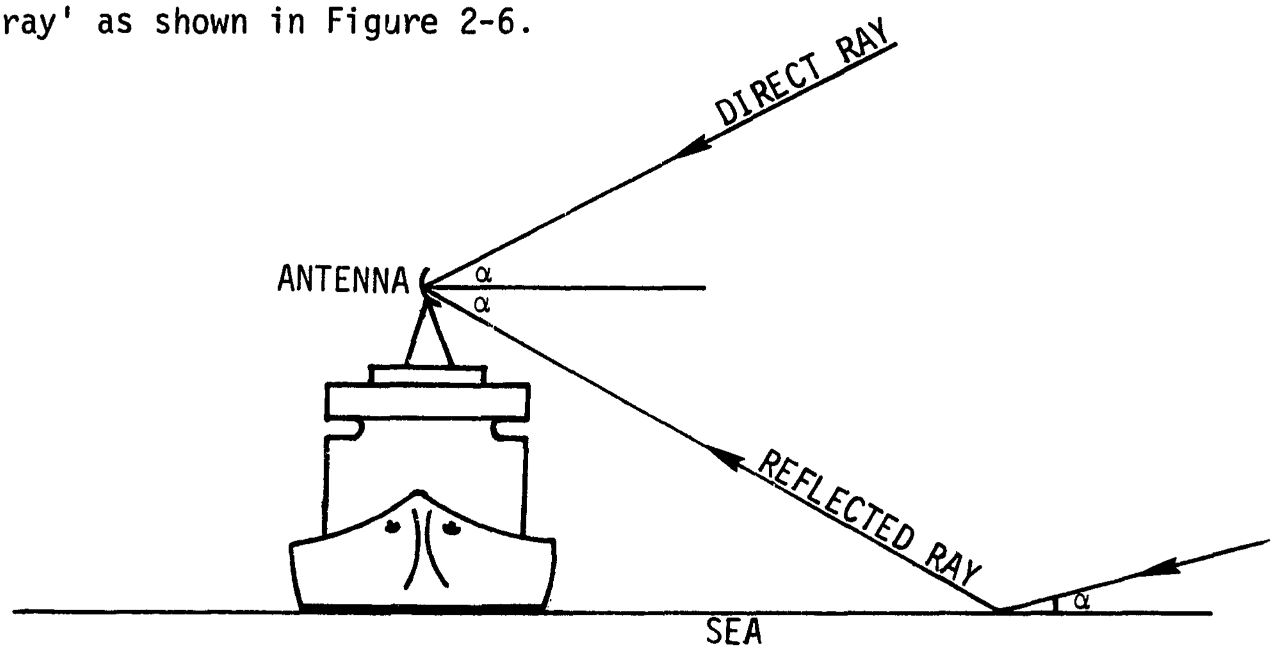

Figure 2-6 MULTIPATH RAY GEOMETRY

Because of the distance to the satellite, the angles at which the two rays arrive are the same. The phase shift caused by reflection, and the phase difference caused by the increased transit time of the reflected ray will cause interference and hence signal fading at the antenna. The magnitude of the effect is dependent on the following factors:

- RF carrier frequency
- Refractive index
- Sea state
- Ship motion
- Satellite elevation angle

- Antenna pattern
- Antenna/Signal polarization

The refractive index of the sea/air boundary is a physical constant and is effectively invariant, as the perturbations caused by temperature variation are negligible.

The sea state is a measure of the roughness of the reflecting surface. The highest amplitude of reflected rays and hence the largest values of multipath fading occur in calm seas. This represents a worst case condition for the design. Sea state is shown as a parameter of the curves in Figures 2-7 and 2-8.

The effect of ship's motion is two-fold. Due to the height of the antenna above the waterline, ship's roll causes changes in the relative path lengths and hence phases of the direct and reflected rays, amplitude modulating the signal at a slow rate. In addition, for antennas without elevation tracking, the effect of ships motion on the antenna pattern is to require that gain be provided at lower elevation angles than would be necessary if the ship did not roll.

The effect of satellite elevation angle is worst at very low angles ($<5^0$), as might be expected, but the effect at larger angles is dependent on the polarization of the signal. This is illustrated in Figures 2-7 and 2-8, from Roberts,[16] for horizontal and vertical polarization respectively.

It is apparent that the amplitude of the reflected signal is considerably less for the vertically polarized signal than for the horizontal polarization. At about 6-7^0 elevation the vertically polarized signal virtually disappears. This angle is the 'Brewster Angle' of the sea/air interface.

Because of the larger value of fading at low elevation angles it would

(16) Roberts, Sam B., Op. Cit.

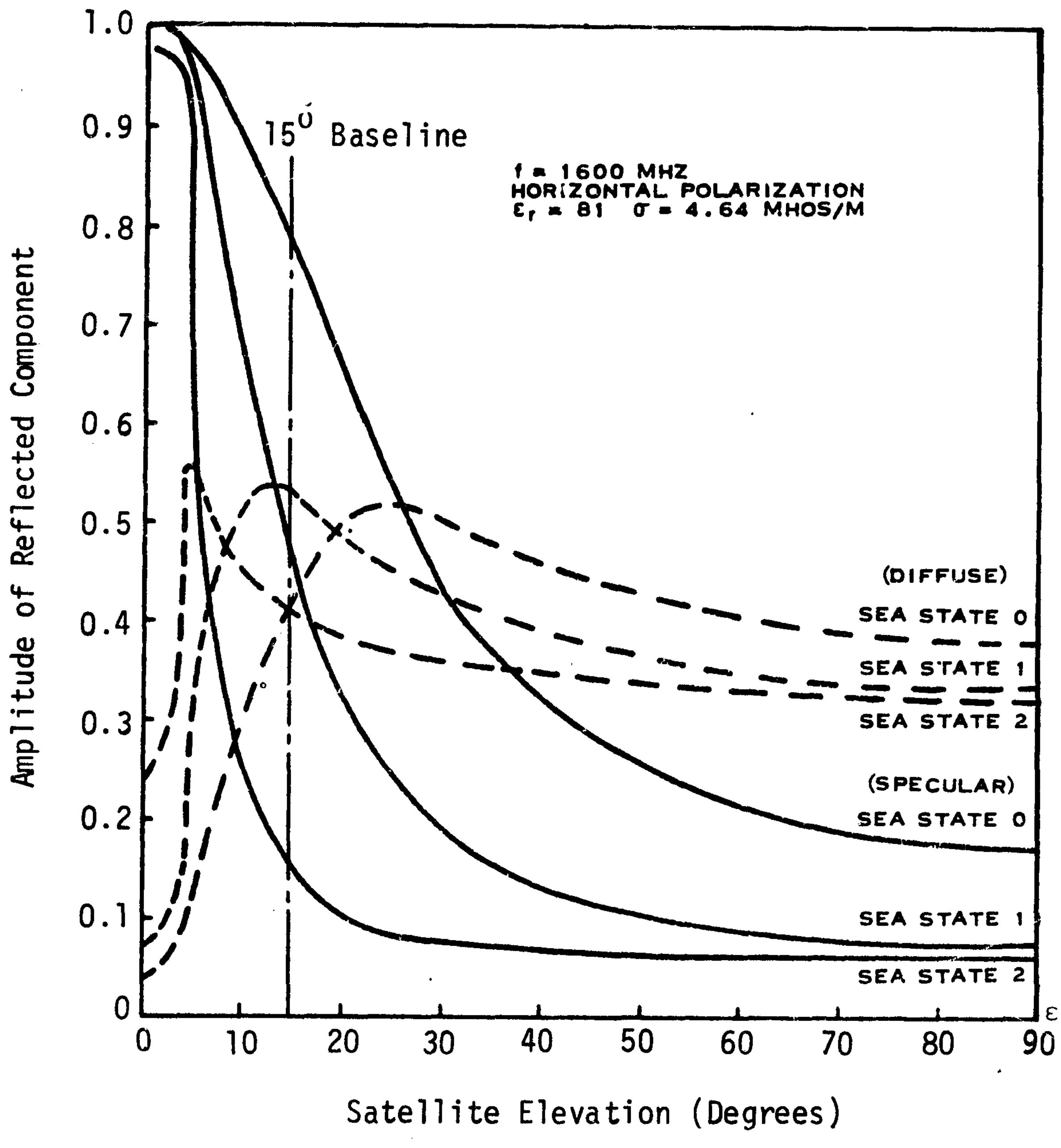

Figure 2-7 AMPLITUDES OF SPECULAR AND DIFFUSE REFLECTED COMPONENTS VERSUS SATELLITE ELEVATION WITH RESPECT TO ANTENNA (HORIZONTAL POLARIZATION)

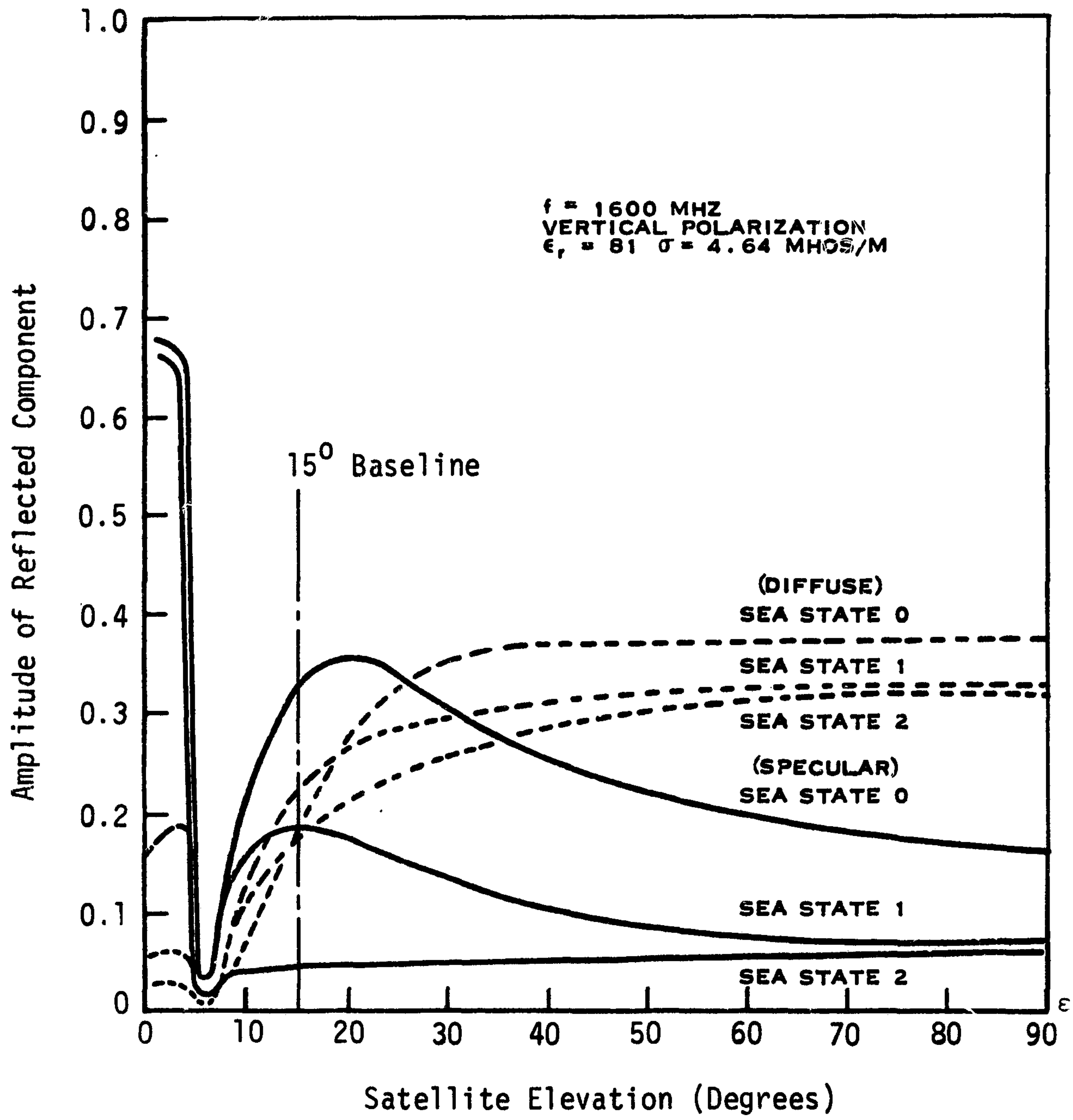

Figure 2-8 AMPLITUDES OF SPECULAR AND DIFFUSE REFLECTED COMPONENTS VERSUS SATELLITE ELEVATION WITH RESPECT TO ANTENNA (VERTICAL POLARIZATION)

obviously be advantageous to design an antenna whose pattern is able to discriminate against the reflected ray at low elevation angles. Because of the effect of ship's motion, it is very difficult to accomplish this without either stabilizing the antenna against ship's motion or providing means for tracking in elevation. Both solutions increase the cost of the installation considerably.

The difference between the curves of Figures 2-7 and 2-8 suggest that the signal/antenna polarization orientation could be used to provide a measure of protection against multipath effects. Examination of the effect on the phase of the reflected signals provides further evidence in favor of such a solution. Figure 2-9, from Skolnik,[22] shows that the phase of the horizontally polarized signal is completely reversed, at all elevation angles. If the incident signal is circularly polarized, then the reflected signal will have its horizontal component reversed, and the vertical component reduced, but with unchanged phase (at all angles above $\sim 10^{\circ}$) and the resultant signal will hence be polarized elliptically, in the opposite sense to the incident signal.

Thus, if the antenna is circularly polarized so as to receive a direct signal of the same sense, it will discriminate against the reflected signal. Because of the ellipticity and the diffusion effects, the isolation will be imperfect. However, a considerable improvement in multipath discrimination can be made by the adoption of circular polarization for both the satellite transmitted signal and the shipboard antenna. For this reason, it can be concluded that circular polarization should be specified for MARSAT system operation.

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(22) Skolnik, M.I. "Introduction to Radar Systems" McGraw-Hill Book Company. 1962.

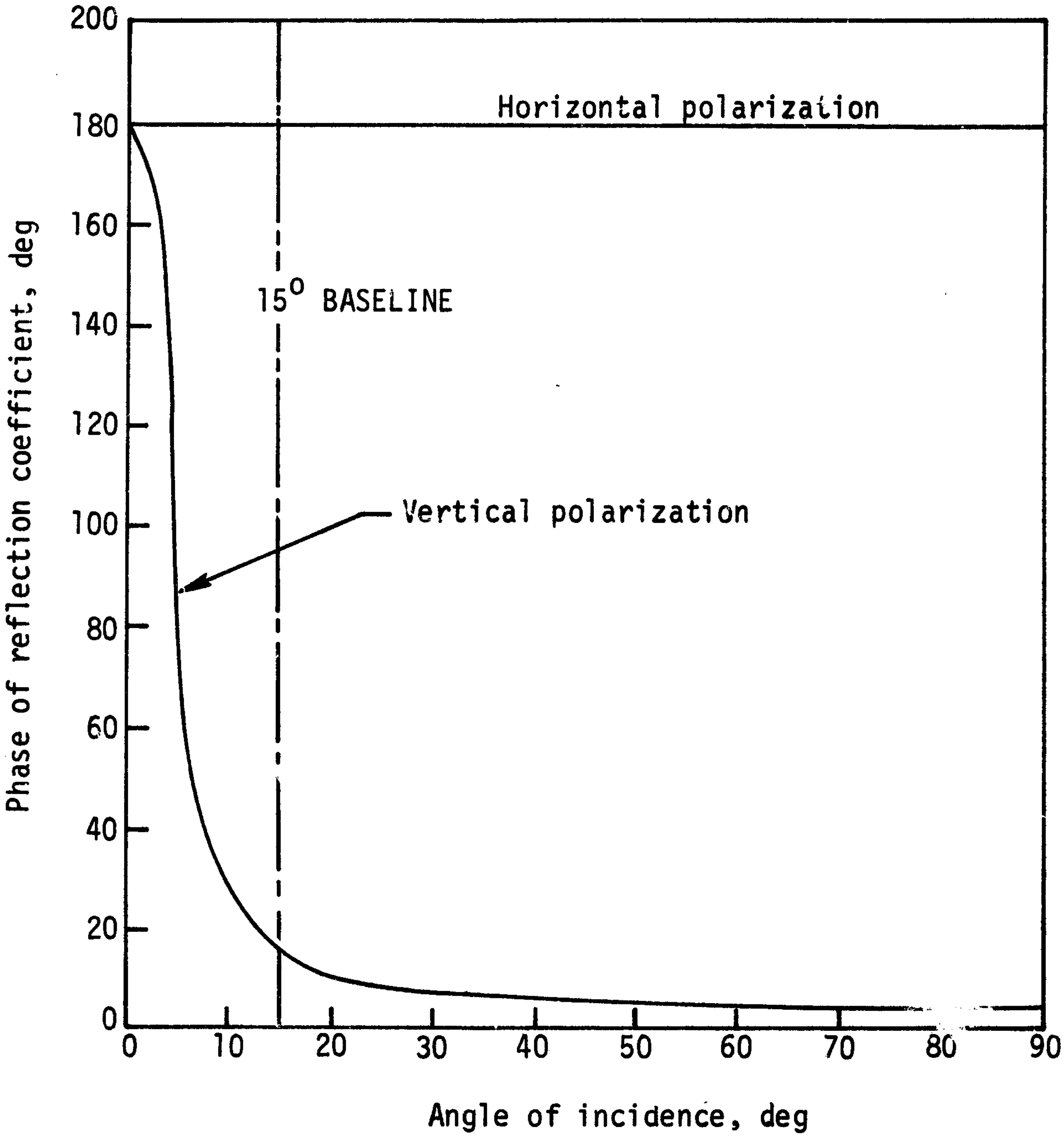

Figure 2-9 PHASE OF REFLECTED SIGNAL VS ELEVATION ANGLE AT 1600 MHz

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2.4 MARITIME ENVIRONMENT

One of the major categories of constraints which affect the design of a small satellite communications terminal for civilian maritime use is the shipboard environment. The potential user vessels spend a great deal of time at sea, and such equipment must be configured to insure reliable operation with relatively infrequent maintenance. The physical factors of the general maritime environment which can affect a shipboard satellite communications antenna include the following:

- Ambient temperature range
- Temperature and humidity cycles (i.e., diurnal and seasonal variations)
- Sunlight and radiant energy variations
- Pressure and wind variations and extremes
- Waves
- Dripping water as encountered in rain
- Wind driven spray (salt water)
- Icing or build-up of ice on exterior surfaces
- Internal and external condensation of moisture
- Ships motion: headway, surge, leeway, sway, heave, roll, pitch and yaw
- Shock transmitted through ship structure
- Existing electromagnetic fields

2.4.1 Temperature, Humidity, Insolation, Rain and Spray

A satcom terminal on a ship will require an antenna mounted on the ship's exterior, typically in a location where it will be exposed to the full force of wind and weather. Various authorities[23,24,25] have specified the extreme conditions encountered at sea for the purpose of establishing design criteria for electronic equipment. Based on this information the following environmental limits are considered reasonable for a satcom antenna installed on a merchant ship which may traverse any of the navigable waters of the world:

1) Temperature: -28°C to 70°C
2) Relative Humidity: 30% to 100%
3) Temperature-Humidity Diurnal Cycle: 25°C to 65°C at 100% Humidity
4) Spray: 2.54 mm diameter drops, 0.3 per cubic cm at 110 km/hr velocity, impinging at any angle
5) Icing: Ice buildup to 15 cm thick and 22 kg/m^2 (4.5 lbs/ft^2) of added weight on exposed surfaces.

The 70°C maximum ambient temperature is higher than the true ambient at sea by about 30°C. This extreme is quoted by the specification writers to allow for absorption of radiant energy directly from the sun. The combination of insolation, ambient temperature and humidity variation on a daily cycle can have an adverse effect on exposed electronic equipment. The diurnal cycle continuing over a period of days constitutes one of the harsher conditions encountered by maritime electronics equipment.

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(23) MIL-E-16400 F (Navy): Military Specification for Electronic Equipment, Naval, Ship and Shore: General Specification; February 1966.

(24) Her Majesty's Stationary Office: Radio for Merchant Ships, The General Post Office Performance Specification. London,1965.

(25) Myers, J.S., Holm, C.H. and McAllister, R.F. "Handbook of Ocean and Underwater Engineering" McGraw-Hill Book Company, New York, 1969.

2.4.2 Shipboard Motion Dynamics and Sea State

Wave motion of water at sea in a storm constitutes one of the most powerful of all existing natural forces. Its effect on a vessel of any size is a prime consideration of all who are concerned with ships,[26-29] from the designers to the crew on-board. Waves affect a satellite communications terminal in the following ways:

- through the ships motion in the waves
- wind driven spray from the waves
- wind driven spray freezing on exterior equipment

Sea waves are primarily the result of the transference of kinetic energy from the wind, which in turn is the result of thermal energy transference in weather phenomena. The friction of the sea surface to the flow of air at their interface results in setting particles of water in motion. These particles pile up, forming ripples or successive ridges on the surface. Continued pressure on the windward side of each ripple or wavelet causes a further depression in the surface and a slight deflection of the air stream at the wavelet crest. This causes a slight lowering of pressure just to leeward of the crest which in turn causes a slight elevation in that area. Thus, the wavelet moves with the wind. With a persistent wind, the wave will continue to grow under this distribution of pressure. The flow of air will continue to supply energy to the wave and it will grow in height and length until the excess energy

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(26) Roll, H.U., "Dimensions of Sea Waves as a Function of Wind Force." Translation by M. St. Denis, SNAME T & R Bulletin, No. 1-19. 1958.

(27) Neuman and Pierson, Principles of Physical Oceanography. Prentice-Hall, Inc. Englewood Cliffs, N.J. 1966.

(28) Gilmer, R., Fundamentals of Construction and Stability of Naval Ships. Second Edition. U.S. Naval Institute. Annapolis, Md. 1959.

(29) Principles of Naval Architecture (Rev. 1967). Edited by J.P. Comstock. Published by SNAME. New York, N.Y.

supplied is entirely consumed in the friction of water molecules in the wave. Energy is supplied as long as the wind velocity exceeds the wave velocity. Height and length of ocean waves are limited not only by this velocity relationship but also by water depth, breaking at the crest, wind force, wind duration and fetch (which is the distance a wave has to run before it is broken up by intervening land).

Ocean waves are very irregular. Figure 2-10 shows a typical record of wave motion past on observation point.

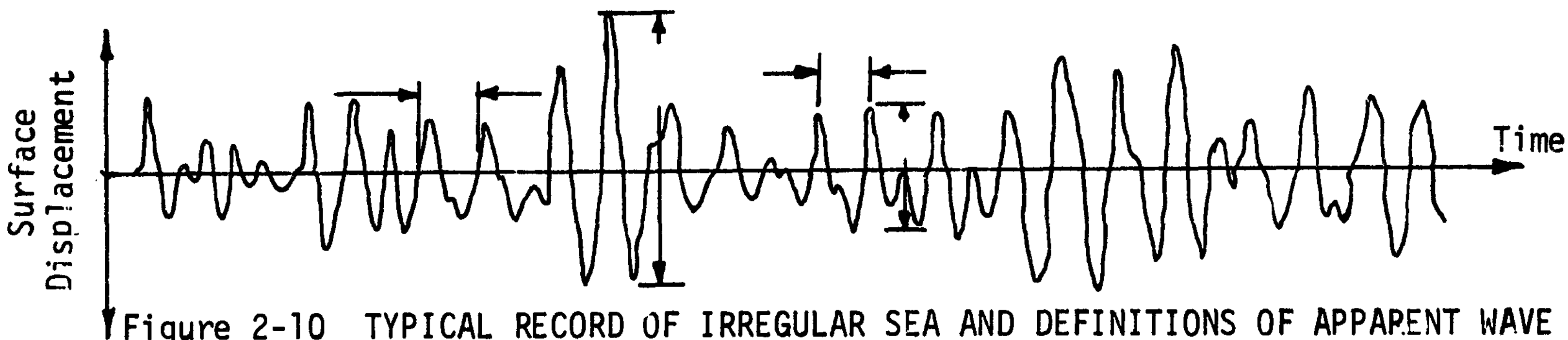

Figure 2-10 TYPICAL RECORD OF IRREGULAR SEA AND DEFINITIONS OF APPARENT WAVE HEIGHTS ($\tilde{h}_w$) and PERIODS ($\tilde{T}$)[29]

Naval architects and oceanographers have applied Fourier type harmonic analysis to the study of ocean waves with comparatively simple and highly useful results. A wave system such as that of Figure 2-10 obviously contains energy at many different harmonically related frequencies. Naval architects typically use a trochoidal wave model, although the characteristics of the motion of a ship in a seaway can be approximated with sufficient accuracy for most purposes using a simple sine wave of a single frequency as an ocean wave model.

To determine the ships motion that can be expected to be generated by waves, for the worst case (e.g., 99%), it is necessary to determine the worst expected sea conditions. The characteristics of waves are defined

(29) Comstock, J.P. Op. Cit.

by velocity, period, height and length. These quantities in regular waves are related to each other and to the wind force.*

Wave heights of 90 ft (27.4 m) have been observed at sea. Heights of 60 ft (18.3 m) have been measured by an instrumented ship.[27] Figures 2-11 and 2-12 from H.U. Roll's analysis of data from North Atlantic weather ships[26] show a smooth curve for average values of wave height and period. Worst case values from the distributions given by Neuman and Pierson[27] are about twice the average. This indicates that a steady state wind of 75 knots (140 km/hr) would build wave heights to about 18 m with periods of 10 to 11 seconds. Therefore, ship's motion generated by 18 m high waves must be considered.

Ship motion at sea can be divided into linear and rotational components. It can be further divided into steady state, uni-directional components (headway and leeway) and transient components, which are oscillatory and irregular in an irregular sea. Headway is simply the average velocity along the ship's longitudinal axis (parallel to the keel, bow direction positive). Leeway is the average velocity (downwind) component at right angles to the keel or the headway component. (To keep plus and minus signs correct for trigonometric functions positive leeway is a set to port). Figure 2-13 shows the transient components of ship motion (with

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* In Reference 29, the following approximate relationships are given for trochoidal wave structures:

$$V = 1.34\sqrt{L}$$
$$T = 0.442\sqrt{L}$$

where V is the velocity in knots, T is the period in seconds and L is the wavelength in feet. Significant wave height (H) (trough to crest) is given in the empirical equation (from Ref. 27).

$$H = 1.82 \times 10^{-2} V^2$$

where V is wind velocity in knots averaged over 6 hours or more and H is in feet.

(27) Neuman and Pierson. Op. Cit.

(26) Roll, H.U., Op. Cit.

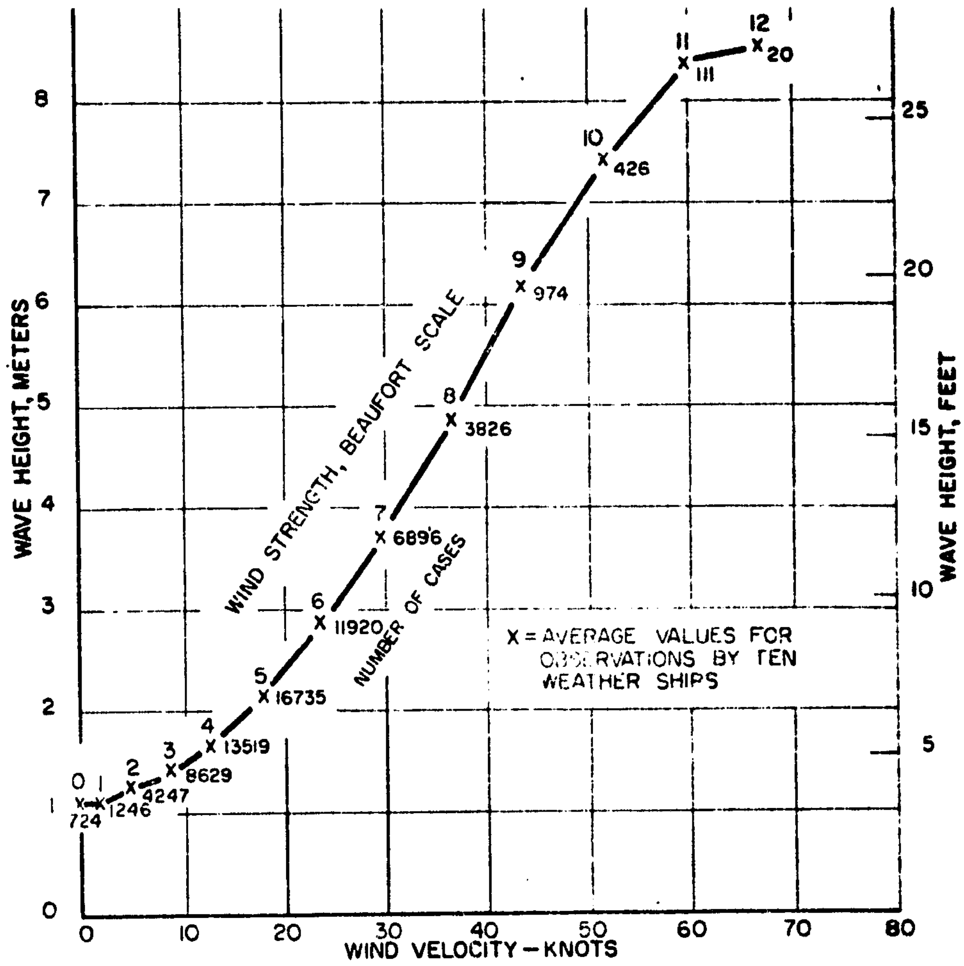

Figure 2-11 WAVE HEIGHT VERSUS WIND VELOCITY

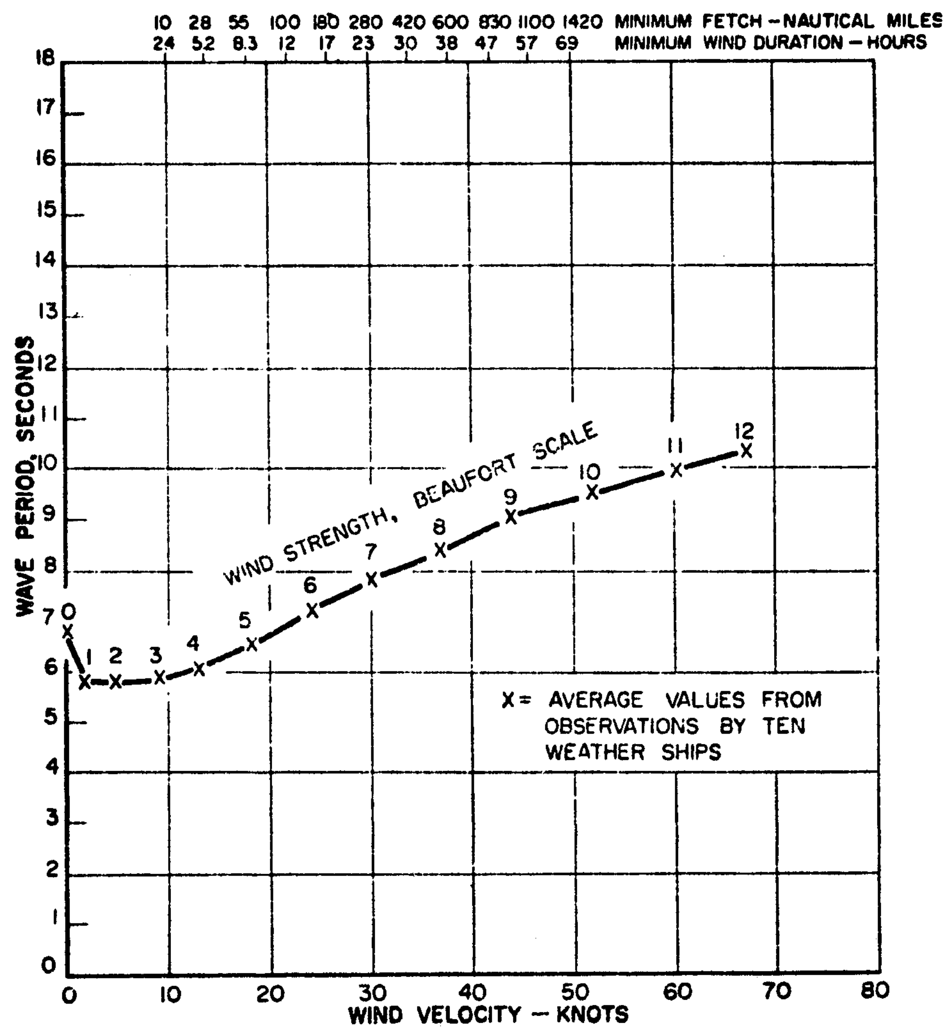

Figure 2-12 WAVE PERIOD VERSUS WIND VELOCITY

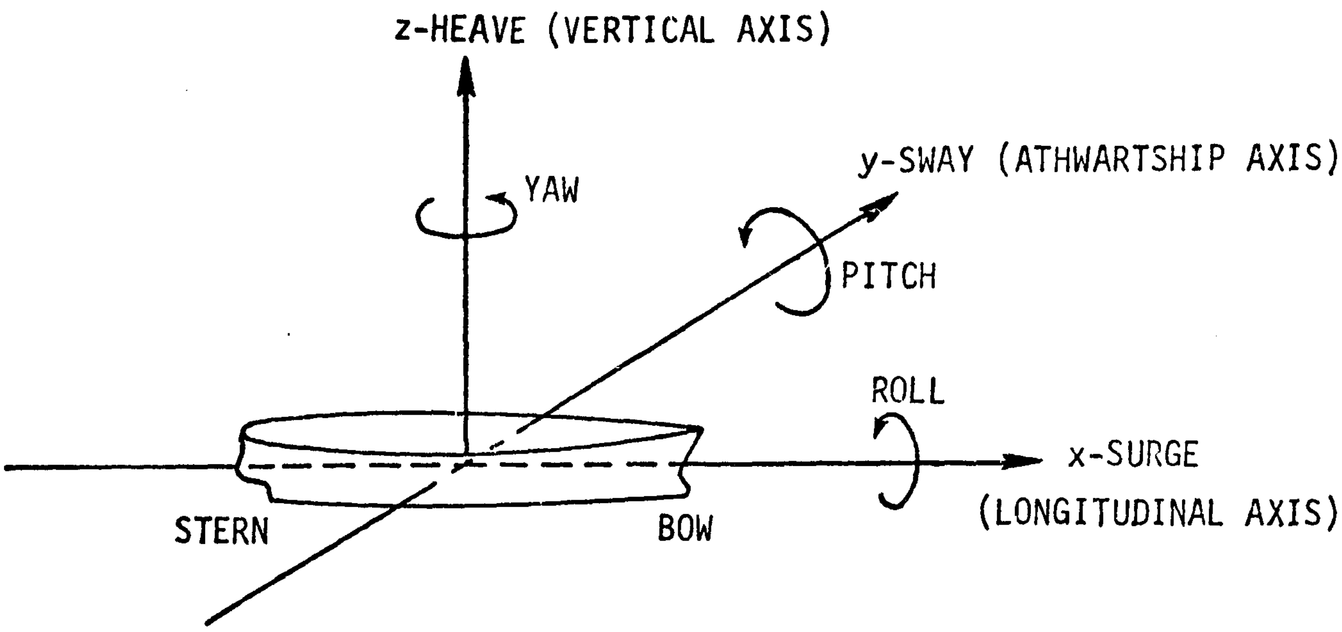

Figure 2-13 TRANSIENT COMPONENTS OF SHIP MOTION

arrows in the positive directions) superimposed on a ship's hull with the bow to the right.

Since a ship may head in any direction and operate at all ocean points in a satellite's coverage area, the shipboard terminal antenna assembly must be capable of transmitting and receiving on all bearings in azimuth and at all elevation angles from horizon to zenith. Aside from its ability to withstand the forces associated with accelerations in all the component directions, surge, heave and sway impose no constraint on a shipboard satcom terminal. Transient displacements along any of the three axes are of the order of one part in 10^6 compared to satellite distance. However, roll, pitch and yaw do impose constraints on the terminal's antenna which must keep the satellite within its beam.

Unless the antenna's beam pattern is hemispherical, plus an additional 20° to 45° to keep the satellite within its beam when the ship rolls, the antenna will have to be trained in azimuth, or in elevation, or both, to keep it pointed toward the satellite within the required tolerance in both coordinates. Tracking ability, including necessary servomechanism responses, depends upon the type of antenna and pattern utilized and upon the characteristics of the ship motion in pitch, roll and yaw.

Rolling is one of the most undesirable characteristics of a ship. As an approximation, if the satcom antenna will meet the constraints imposed upon it for rolling under all conditions, it will meet those imposed by yawing and pitching. Given the wind conditions, and/or sea conditions, course and speed and characteristics such as metacentric height,* beam, moment of inertia, damping coefficient, etc., it is possible to predict the maximum expected roll.** Some of these characteristics are difficult to determine and maximum roll is normally derived by consideration of stability characteristics. In practice, the rolling characteristics of a ship vary greatly. The period of natural

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* Metacentric height is the distance between the ship's center of gravity and the intersection of the vertical through the center of bouyancy, when the ship is keeled slightly, and a perpendicular to the keel through the center of gravity.

** Gilmer[28] indicates that the maximum amplitude of synchronous roll (Φ_r) is a combination of the two equations:

$$\frac{\Phi_r}{\alpha_M} = \frac{2\pi D}{W_L A}\left(\frac{GM}{g}\right)^{\frac{1}{2}} \approx 10, \text{ and } \Phi_r = \frac{\pi}{W_L}\left(\frac{3\pi D \alpha_M}{2 \beta g}\right)^{\frac{1}{2}}$$

where: α_M = maximum wave slope = $\pi H/W_L$, H is the wave height.
A = coefficient of added moment of inertia due to bilge keels
D = displacement
W_L = wave length
β = damping coefficient

Since α_M may be as high as 9° it is evident that synchronous rolling is dangerous.

roll depends to a first approximation on the metacentric height and the beam of the ship. The amplitude depends mainly on three items:

- the ratio of the wave train period (T_W) passing the ship to the natural period (T_Φ) of the ship's roll
- Damping due to bilge keels, fin stabilizers or anti-rolling tanks
- The effective height of the waves as a driving force.

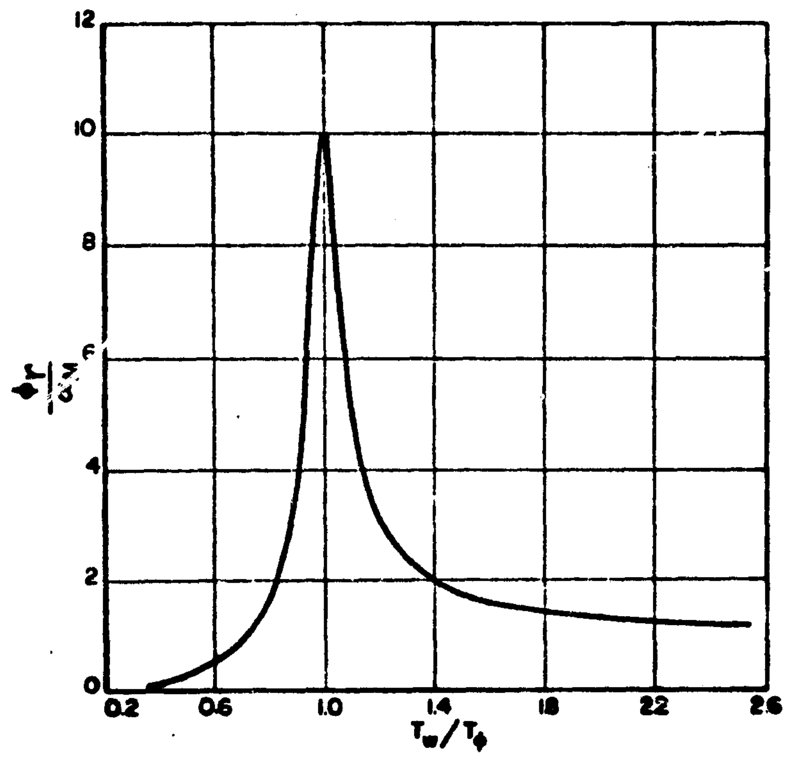

Figure 2-14 AMPLITUDE OF FORCED ROLLING FOR VARIOUS VALUES OF T_W/T_Φ

As shown in Figure 2-14, when synchronism exists ($T_W = T_\Phi$), the amplitude of roll (Φ_r) of a typical ship is 10 times as great as the maximum wave slope, α_M, and approximately 10 times as large as when the roll is completely unsynchronized ($T_W/T_\Phi \geq 3$). The metacentric height of a vessel, such as a tanker, will vary greatly between loaded and ballasted condition. Since large tankers travel in ballast about half the time the worst of the two conditions must be considered. This is the ballasted condition. Rolls up to ±30° amplitude have been reported by 1000 foot long supertankers for periods of several hours at a time when proceeding in ballast. Smaller ships with greater metacentric heights and narrower beams tend to roll to greater amplitudes. Roll amplitudes up to 50° port and starboard have been reported by 300 ft. long Coast Guard cutters.

The roll amplitudes reported are probably greater than the actual roll. They are generally measured by a pendulum type inclinometer mounted on the bridge of the ship 9 to 24 meters above the ship's center of gravity. (A ship rolls about an axis which coincides approximately with the center of gravity.[29]) As shown in Section 3.4, the overshoot of an undamped pendulum a distance L above the roll axis is proportional to L. The overshoot error may be equal to the actual angle of roll. In the case of a large ship in ballast with a roll period of about 12 seconds and an undamped pendulum inclinometer on the bridge 24 meters above the center of gravity, the inclinometer error may be

$$\Delta\phi = \frac{L\,\omega_1^2}{g}\phi_o = \frac{24 \times (2\pi/12)^2}{9.8}\phi_o = 0.62\,\phi_o$$

where ω_1 is the angular frequency of roll, g is the acceleration due to gravity and ϕ_o is the true roll amplitude. In this case, an indicated roll of 30° would be a true roll of 30°/1.62 = 18.5°. Similarly, a Coast Guard cutter with bridge height of about 12 meters and the same roll period will have approximately 31% overshoot. An indicated roll of 50° is thus a true roll of about 38°. These values have been noted to

(29) Comstock, J.P., Op. Cit. pg. 670.

agree closely with those given in a recent paper submitted by the Norwegian delegation to the Special Joint Meeting of the C.C.I.R. in February 1971 in Geneva.(30) Figure 2-15 is from this report, and shows the expected roll amplitude versus ship size and speed.

For purposes of determining the constraints that rolling places on satcom antennas, two cases will be considered as representative. First is the small vessel of about 50 meters in length, which has a maximum expected roll amplitude of ±45°* and second is the large vessel, 150 meters in length and over, which has a maximum expected roll amplitude of ±30°.

The period τ_r of a ship's roll is defined as the time in seconds for the ship to roll from an upright position to maximum amplitude on one side, return, pass through the vertical position to maximum amplitude on the other side and finally to return to the initial vertical position. A ship's roll is a harmonic function. The ship acts as a pendulum. It has a natural period of roll. As noted previously, the rolls of maximum amplitude occur when the natural period is the same as the wave period causing the roll. Natural roll period depends upon a number of factors including

- Metacentric height
- Radius of gyration of entrapped water (whether bilge keels are installed)
- Whether a stabilization system such as fins or flume tanks are installed.

As a rough approximation, the natural roll period is given by(28)

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(30) C.C.I.R. document M/241-E. "Systems Providing Radio Communications and/or Radio-Determination Using Satellite Techniques for Aircraft and/or Ships - - The Motion of the Ship Due to the Sea." Norway 28 January 1971

* This corresponds to the inclination test of Mil-E-16400 (Ref. 23)

(28) Gilmer. Op. Cit.

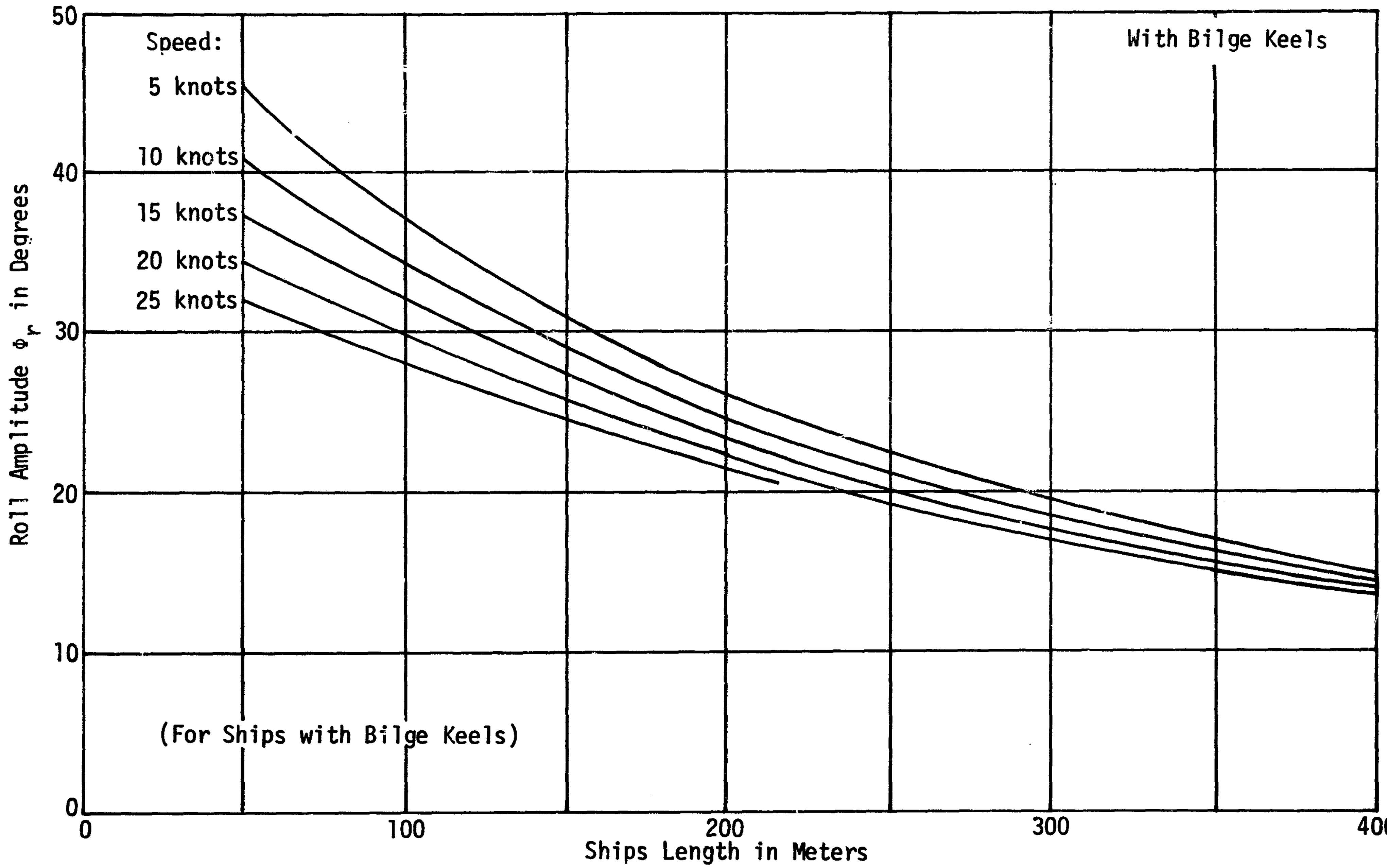


Figure 2-15 LARGEST EXPECTED ROLL ANGLE AS A FUNCTION OF THE LENGTH AND SPEED OF THE SHIP

$$\tau_r = \frac{0.44\ B}{\sqrt{GM}} \tag{2-4}$$

where B is the extreme beam in feet and GM is the metacentric height in feet. A supertanker in ballast may have a GM of 30 to 40 ft (10 m) and a beam of 150 to 170 ft (50 m). Under these conditions, $\tau_r \approx 12$ seconds. Under loaded conditions, the GM decreases to approximately 10 feet and $\tau_r \approx 22$ seconds. Figure 2-16 indicates average roll periods.(30) Using the same ship lengths as considered above for roll amplitude, the small vessel roll period is rounded-off to a value of 6.28 (2π) seconds and the large ship roll period is taken as 12.56 (4π) seconds, for ease of computation. Although the rolling of a ship is irregular, the worst condition for stabilization of the satcom antenna would be that of synchronous rolling at the vessel's natural period of roll. Under these conditions, the roll will be sinusoidal and the amplitude of roll may be ten times as great as when the roll is not synchronized with the wave period. For purposes of this investigation, the following worst conditions will be used.

Small ship - Under 150 meters in length

Roll amplitude (Φ_r) $\leq$ 45°

Roll period (τ_r) $\geq$ 6.28 seconds

Large ship - Over 150 meters in length

Roll amplitude (Φ_r) $\leq$ 30°

Roll period (τ_r) $\geq$ 12.56 seconds

Using the usual definitions of satellite azimuth and elevation, it is obvious that a rolling ship will impose some requirement for antenna motion. If, for example, a ship is on a steady course, with no yawing, rolling or pitching, the satcom antenna may be trained to a bearing angle, Θ, measured from the ship's bow and elevated on that bearing through a vertical angle, ϕ, measured from the plane of the ship deck

(30) CCIR Doc. M/241-E. Op. Cit.

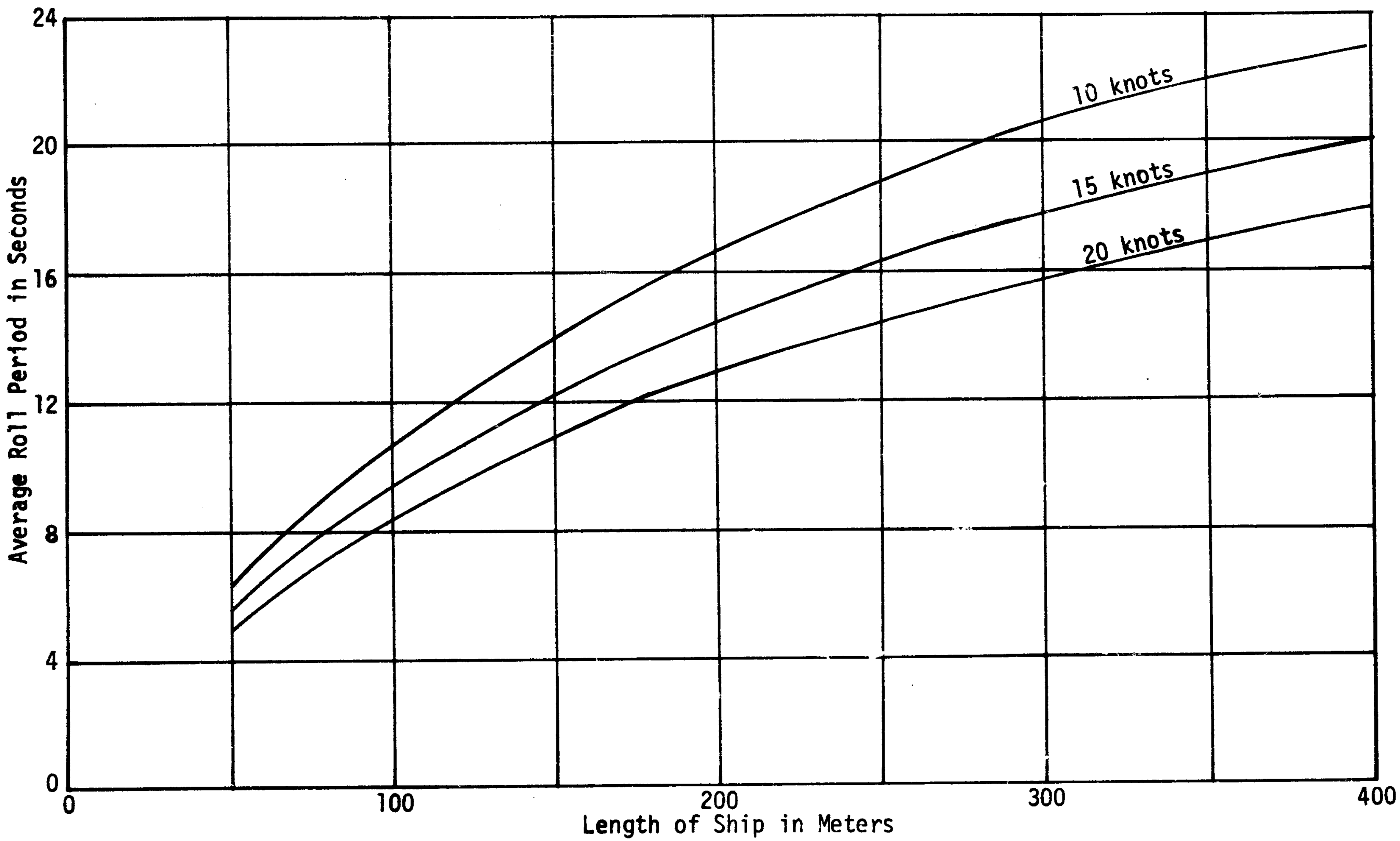


Figure 2-16 AVERAGE ROLL PERIOD AS A FUNCTION OF THE LENGTH AND SPEED OF THE SHIP

so that it is pointed directly at the satellite. When the ship rolls through an angle Φ_r both Θ and ϕ will change to Θ' and ϕ' as shown in Figure 2-17. The amount of change depends upon the original Θ and ϕ as well as Φ. Using the projected angles β and α as defined in Figure 2-17 and letting $\Theta' = \Theta + \Sigma$, where Σ is the azimuth error, it may be seen that

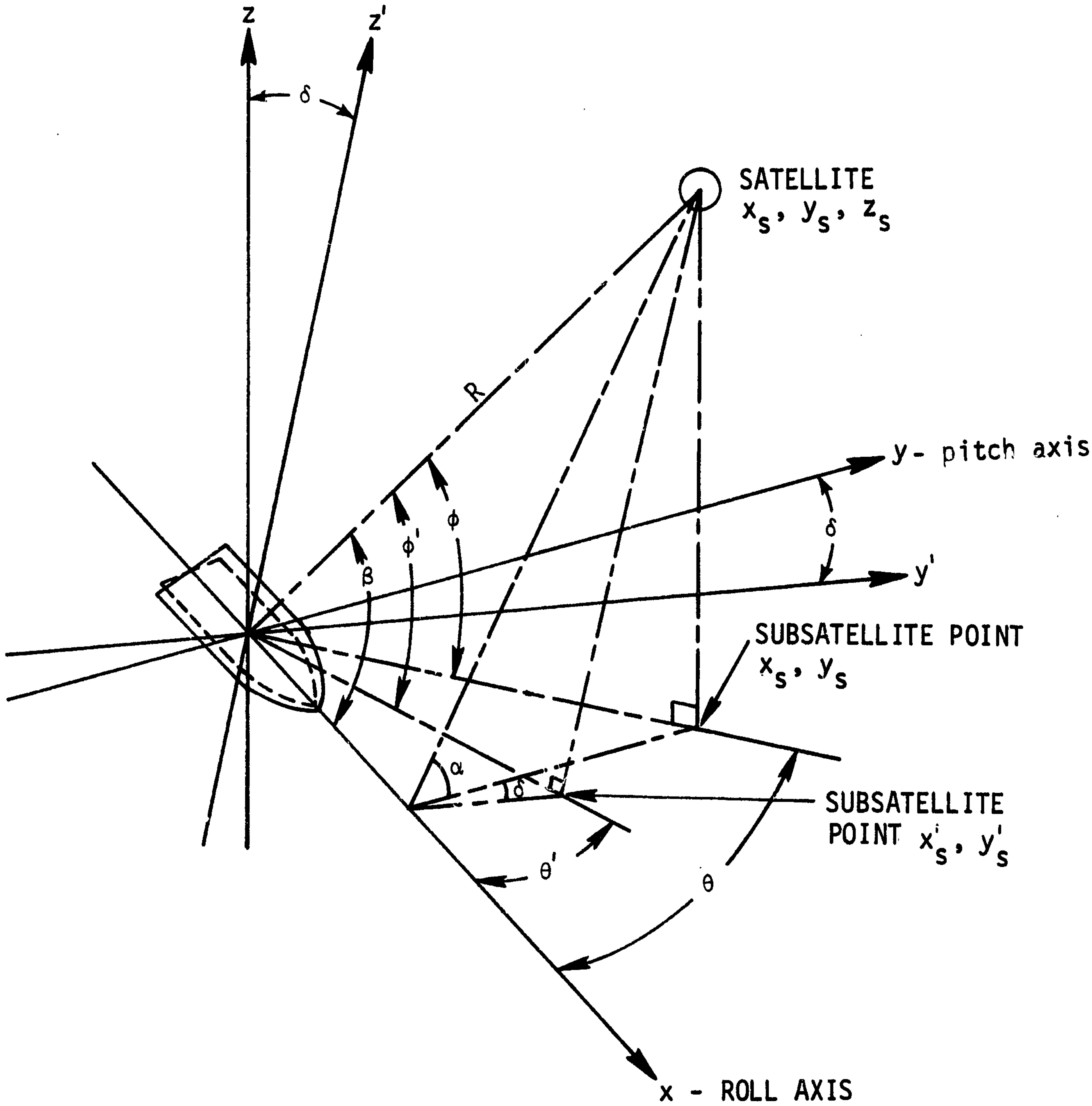
$$\tan \beta = \frac{\tan \Theta}{\tan \alpha} = \frac{\tan (\Theta + \Sigma)}{\cos (\alpha + \Phi_r)} \qquad (2\text{-}5)$$

Since $\tan \alpha = \tan \phi / \sin \Theta$, the azimuth error, when the deck plane is the reference, (i.e., the antenna is trained only relative to the deck plane) is

$$\Sigma = \tan^{-1} \left(\tan \Theta \cos \Phi_r - \frac{\tan \phi \sin \Phi_r}{\cos \Theta} \right) - \Theta \qquad (2\text{-}6)$$

Figure 2-18 is a plot of this equation when the roll is 30° to either side for various elevation angles (ϕ). This corresponds to the worst case roll for a large ship. Similar curves can be plotted for the small ship with roll amplitude of 45°. These curves show that the instantaneous azimuth errors caused by rolling are excessive when the satellite elevation is high. The limiting case occurs when $\Phi_r + \phi = 90°$ which, when $\Theta = 90°$ or 270° means that the satellite is at the zenith when the ship reaches maximum roll amplitude. Under this condition the azimuth error, Σ, approaches 90° when the bearing is near 90° (or 270°). When $\Phi_r + \phi > 90°$, the azimuth error is 180° for $\Theta = 90°$ (or 270°).

For a given elevation angle, ϕ, as the ship rolls from $+\Phi_r$ to $-\Phi_r$, port to starboard, the error, Σ, goes from a positive maximum to a negative. When the ship rolls toward the satellite, the error is greater than when it rolls away from the satellite. Figure 2-19, a



x - y	plane - horizontal
x'-y'	deck plane when vessel has rolled its deck plane through an angle δ
ϕ	is elevation angle measured from deck plane when it is coincident with the horizon plane
ϕ '	is elevation angle measured from deck when rolled to angle δ from the horizontal
θ	is azimuth angle measured from bow when deck plane is horizontal
θ '	is azimuth angle in deck plane when rolled to angle δ

Figure 2-17 EFFECT OF SHIPS ROLLING ON ELEVATION AND AZIMUTH

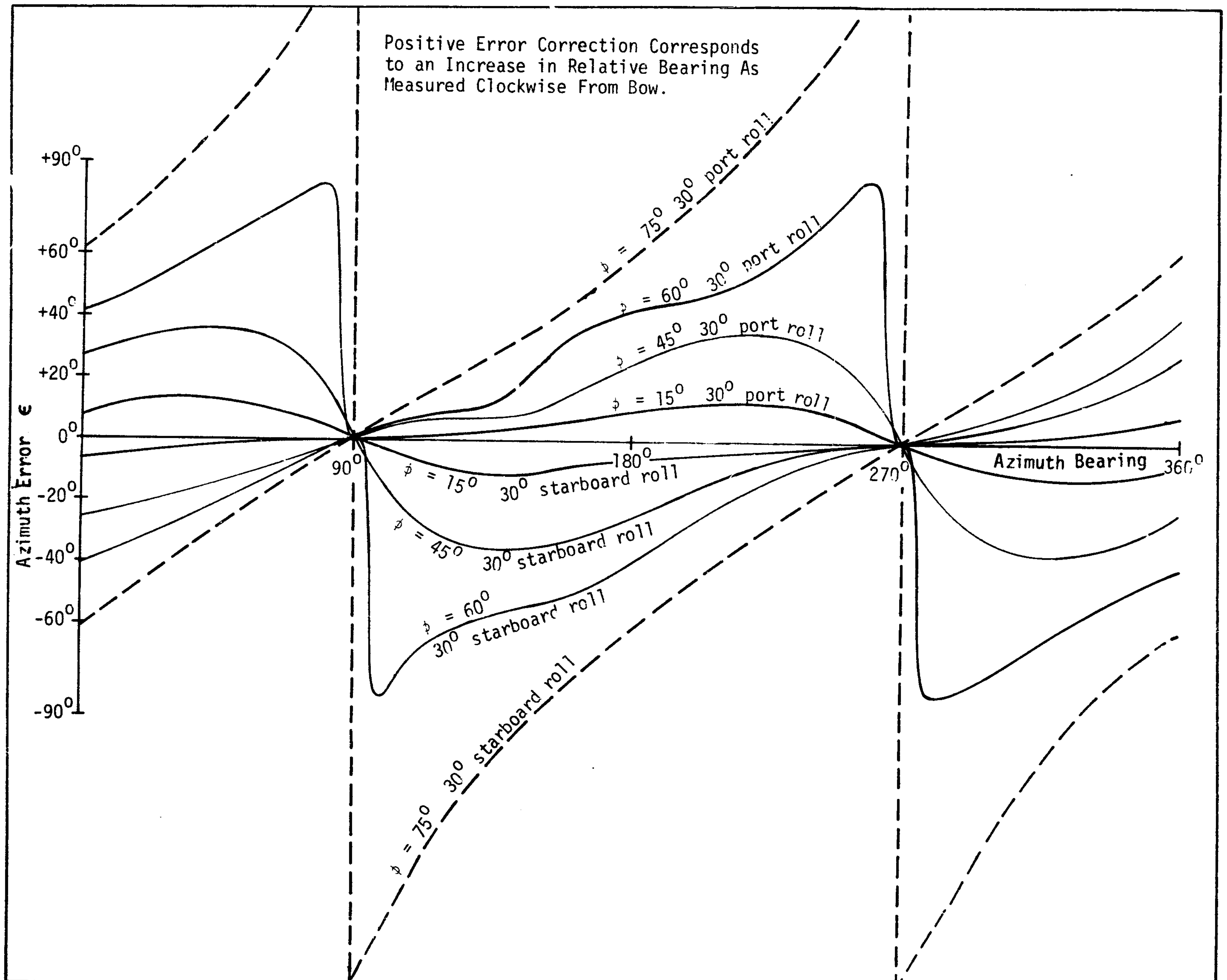


Figure 2-18 MAXIMUM AZIMUTH ERROR (ε) FOR 30° ROLLING (Φ_r=30°) AS A FUNCTION OF ELEVATION (Φ) AND AZIMUTH (Θ). ANTENNA STABILIZED TO DECK PLANE ONLY.

plot of Equation 2-6 for ϕ = 35.3° and Θ = 45°, illustrates this point. Thus, the potential azimuth error, or change relative to the deck plane, places a severe constraint on an antenna which is to track in azimuth only, when the satellite elevation ϕ exceeds 90° minus the maximum roll amplitude and the satellite direction is near the beam of the ship. The antenna servo system must almost instantaneously drive the antenna through an arc of 90° to reach the satellite bearing. This constraint indicates the need for horizontal stabilization (i.e., elevation angle tracking) of a satcom antenna such that the azimuth error caused by roll is minimized.

However, if the assumed fan beam has a very large elevation beamwidth, such as 90° + $2\Phi_r$, then the instantaneous change in relative azimuth of 180° (from 90° to 270°) is only mathematical, in that the antenna beam would still contain the satellite. Figure 2-18 must be interpreted with this facet in mind. For a very large fan beam, e.g., 150 degrees in elevation for large ships and 180° for small ships, the largest azimuthal error occurs when the relative bearing of the satellite is near the bow-stern line of the ship.

In summary, the constraint rolling imposes on an antenna in <u>elevation angle</u> is worst when the relative satellite bearing is 90° (or 270°). On these bearings, the amplitude of the change in elevation is essentially equal to the amplitude of the roll. If the vessel is rolling to an amplitude of 45°, the total change in elevation is ±45°, i.e., 90°.

Angular velocities and accelerations depend upon the effect of the irregular sea waves which can be approximated quite well by sinusoids. The dynamic constraints imposed by ship's roll on the satcom antenna beam pointing subsystem may be summarized as in Table 2-4 for the two representative ship size categories.

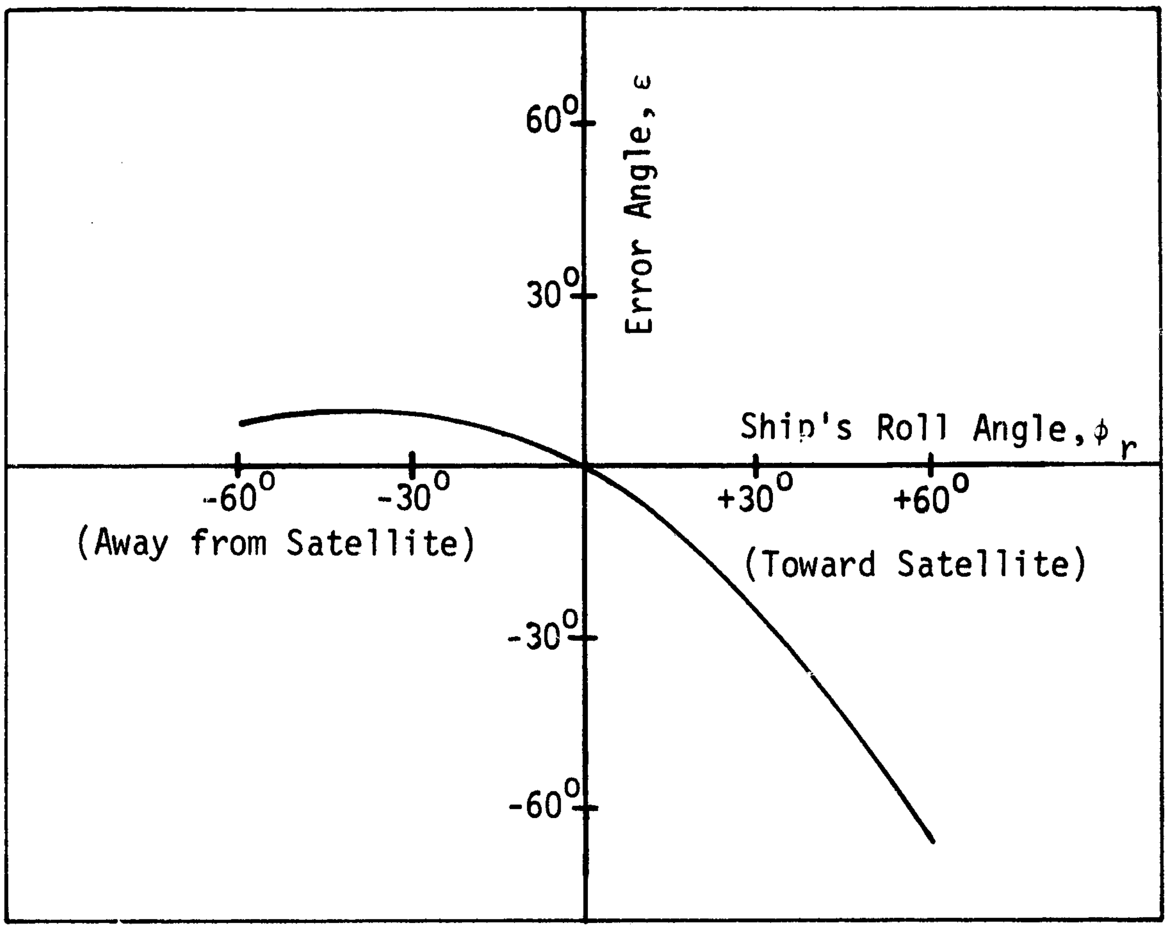


Figure 2-19 AZIMUTH ERROR AS SHIP ROLLS FOR RELATIVE SATELLITE BEARING, Θ = 45° AND ELEVATION ϕ = 35.3°

TABLE 2-4
SHIP'S ROLL RATES

Vessel Type	Maximum Roll Amplitude	Roll Period	Maximum Roll Rate	Maximum Angular Acceleration
Small Ship (<150 meters long)	~ 45°	6.3 sec.	~45°/sec	~ 4.5°/sec^2
Large Ship (>150 meters long)	~ 30°	12.6 sec.	~15°/sec	~ 7.5°/sec^2

2.4.3 Wind, Vibration, Corrosion

Wind velocities at sea vary from zero to more than 240 km/hr. The high velocity range occurs in hurricanes or typhoons. (For example, a velocity of 129 knots was noted in a hurricane near Puerto Rico about 10 years ago.)[31] Wind velocities are not steady. Gusts lasting for short time intervals may register a velocity 30 to 50% higher than the average. The highest wind velocities encountered in extra tropical cyclones seldom exceed 75 knots. In fact, 75 knots is the design criterion used by the U.S. Navy,[23] and is considered above the threshold for hurricane classification. H. U. Roll of Holland made a statistical analysis of wind and waves as reported by the North Atlantic weather ships.[26] He reports a lower velocity encountered 99% of the time. In 65,976 observations over a two-year period, 65,419 or 99.15% were 55 knots or less. Thus, 75 knots for wind velocity is a fairly safe design criteria for almost all satcom terminal operations. This implies that the antenna should be designed to keep the satellite within its beam with a modest, allowable degradation of performance (e.g., 1 dB of gain), with wind pressures up to about 150 kg/m^2 for flat surfaces. Operation in hurricanes is also important for emergency communication. Hence, performance should degrade gracefully at velocities greater than 75 knots. For example, adequate pointing accuracy and antenna gain should be available for at least telegraphy access and exchange.

Examination of existing electronics specifications and information from potential MARSAT users who operate fleets of merchant vessels indicates

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(31) Reference Data for Radio Engineers - Fifth Edition. Howard Sams, Inc. New York, 1968. pg. 41-3.

(23) MIL-E-16400. Op. Cit.

(26) Roll, H.U., Op. Cit.

that shock and vibration encountered on ships are generally a constraint on equipment design. The worst cause of shock is "bow slamming." As the ship proceeds at normal speeds in a seaway it can and does occasionally encounter a condition where the length between the crests of the waves is such that the ship will ride up on a wave so that the forefoot of the bow comes out of the water. With some ships, as much as a third of their lengths may clear the water. As the ship passes over the wave in this condition, the bow slams down in the trough with a shock that is as high as 50 times the force of gravity (50 g's) and of several milliseconds duration.

Vibration is caused mainly by propeller tip rotation. It is of low frequency, from 2 to 3 Hz up to about 35 Hz. Some ships transmit these vibrations quite readily throughout the hull and superstructure setting up nodes and anti-nodes. At locations such as the upper portion of the mast, where mechanical resonance may exist, the amplitude of vibration of the ships structure may reach 0.15 cm (0.06 in.) according to surveys made by one fleet operator. The frequency where this maximum occurs if 6-9 Hz.

2.4.4 Limitations in Antenna Locations

Large modern ships place some constraints on the location of satcom antennas. Figures 2-20, 2-21 and 2-22 are the elevations of three examples of modern ships of differing types. While king posts and masts are available away from the bridge deckhouse area, they generally represent a long expensive cable run between antenna and radio room. It is preferable to install the antenna in the deckhouse area. The locations available are the masthead or the wheelhouse top. In the latter location, a clear unobstructed area for all azimuths from horizon to zenith cannot usually be found and two antennas may be necessary. Hence, the masthead is the optimum location.

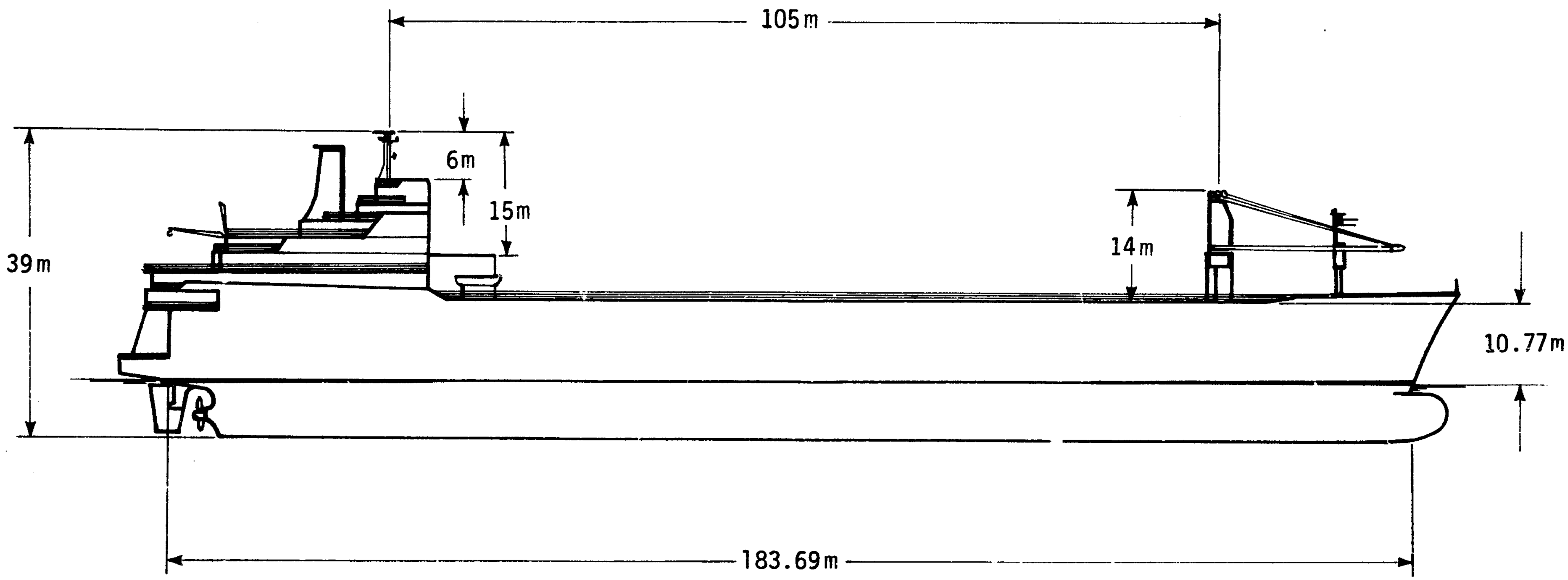

Figure 2-20 EXAMPLE OF A 20,500 DWT CONTAINER SHIP

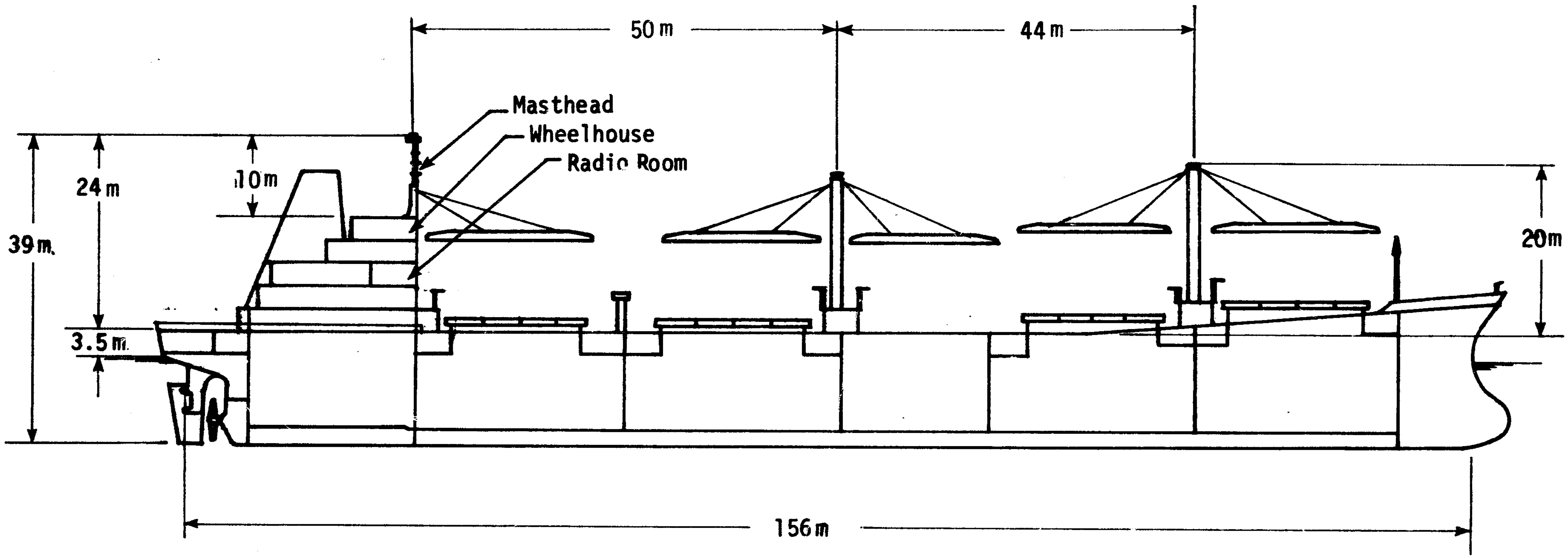

Figure 2-21 EXAMPLE OF AN ORE AND BULK CARRIER 21,500 DWT

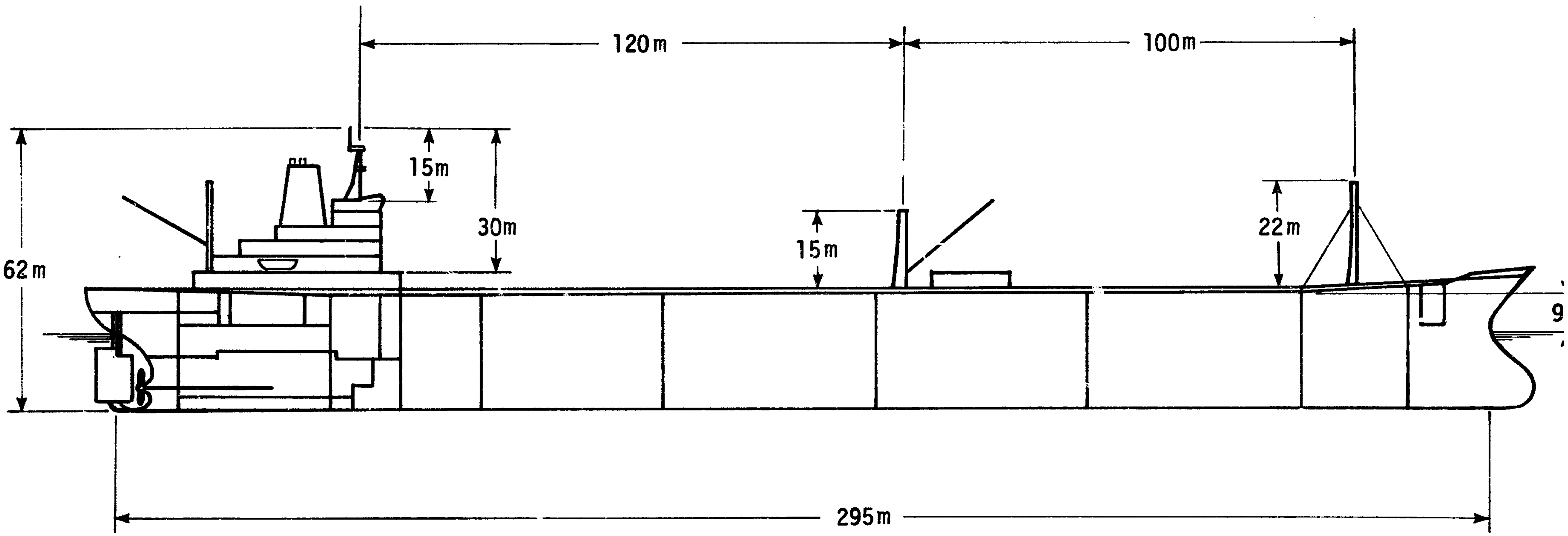

Figure 2-22 ELEVATION OF A 215,000 DEADWEIGHT TON TANKER

While merchant ship masts are relatively uncluttered, the masthead is typically occupied by one or two radar antennas (currently). The radar antenna is usually a 2 to 3 meter long slotted waveguide array which, if mounted forward of a 23 cm mast will suffer relatively little attenuation when pointed aft. Hence, the relocation of a radar antenna accordingly, and installation of a satcom antenna at the masthead, might be considered to be the more practical and least expensive of alternative antenna locations.

2.4.5 Electromagnetic Interference

There are a number of sources of electromagnetic radiation on ships of all sizes and descriptions. Some radiation is noise and some is comprised of signals used for purposes other than communications by satellite. On merchant ships, desired radiation is from radio transmitters in the MF, HF and VHF bands. There is very little equipment on merchant ships emitting in the UHF band which will be utilized for a satcom terminal. Merchant ship radars operate at about 3 and 9 GHz. With equipment meeting CCIR and FCC standards, the likelihood of direct interference from radio equipment is small (unless a harmonic of a transmitting frequency of a VHF-FM transmitter in the marine band happens to fall precisely on the receiving frequency of the satcom terminal in the UHF band at 1540 MHz, which is unlikely).

Noise from electric motors and other electrical equipment is always a possibility on-board ship. This noise is broad-band in character, such as results from an arc caused by discharge across a gap where a voltage potential exists. It can reach high levels in the HF band but is generally attenuated appreciably at UHF frequencies.

Another troublesome source of electromagnetic noise is arc discharge across insulators on HF and MF transmitting antennas. Shipboard antennas are electrically short. When they are energized, high potentials exist across their insulators. When moisture and salt coat these insulators, voltage breakdown and broad-band arcing noise result. Here also, most of the energy is concentrated in the MF and HF bands, but some noise can result in the VHF and UHF bands.

Rigging, such as mast stays and wire rope life lines can also cause interference. This may be broad-band or on discrete narrow frequency bands. In a strong RF field, as generated by one or more HF transmitting antennas, current is induced in the rigging. Unless bonding of wire stays and life lines to the steel hull is in good condition, there is likely to be one or more corroded connections between the wire rope, eyes and the mast, stanchion or deck. These corroded connections act as non-linear electrical elements. They generate harmonic and intermodulation product voltages which send current over the wire ropes on a new family of frequencies. These currents cause radiation at the intermodulation and harmonic frequencies. As in the previous cases, the greatest resulting radiation fields are at frequencies below the UHF band.

Some satcom terminal noise and interference may result from the sources described above. The magnitude of the noise in the UHF band has not been measured on any civilian or merchant ships. The noise sources described can usually be controlled and the noise held to a minimum. Hence, the satcom terminal is subject to only a minimal noise environment on a "clean" ship. The prognosis is that noise from the sources described can be limited to levels which will not cause harmful interference.

2.5 SYMMETRICAL VS NON-SYMMETRICAL ANTENNA BEAMS

Since the beamwidths of the antennas being considered are not narrow, and the major source of relative motion is only that of ships roll, non-symmetrical beams can be considered in which the necessary directivity is only in one axis. For example, a fan beam can be considered in which the elevation beamwidth accommodates all necessary roll angles and the relatively narrow azimuth beam is made to follow the satellite direction by conventional signal tracking or slaving to a shipboard reference, i.e., a compass. Alternatively, a fan beam can be used which covers all azimuth angles, i.e., with a 360° horizontal beamwidth that has a narrow elevation beamwidth which is variable in position.

In general, practical shipboard satcom antenna beams are constrained primarily by ships motion and the manner in which they are pointed. Table 2-5 shows certain constraints imposed on the beam pattern by ships motion for given schemes used to train and elevate or stabilize it. It does not account for long term motion such as from one day to the next. The beam shapes defined in Table 2.5 result from the considerations of Paragraph 2.4.2. In the case of the fixed beam antenna, when the satellite is on a relative bearing of 90° (270°) and the ship rolls away from the satellite, the beam must extend below the angle of the deck plane to an angle sufficient to keep the satellite illuminated at the minimum satellite elevation which is considered operational. For example, if the minimum elevation for a given service quality is 5° above the horizon, the typical large ship, with a maximum expected roll amplitude of 30°, needs vertical coverage from -25° to +90° on all azimuths.

The azimuth errors shown in Figure 2-18 are for a ship on a steady course with no azimuth error introduced by yaw. One way of eliminating yaw error is to slave the antenna to an on-board gyro-compass (north

TABLE 2-5

GENERAL CONSTRAINTS ON BEAMWIDTH DUE TO SHIPS MOTION

Type Antenna	Stabilization Reference	Type Beam	Required Horizontal Beamwidth	Required Vertical Beamwidth
Fixed	None	Hemispherical	360°	minimum operational elev. less roll amplitude to + 90° in elevation
Automatically trained in Azimuth, Manually elevated	North reference only	Wide Eliptical cone	180° cos $(90° - \Phi_{rm})$	2 x maximum roll amplitude
Automatically trained in Azimuth, Manually elevated	Maximum received signal strength	Fan	2 x signal sensitivity to bearing change + maximum lag in servo followup	2 x maximum roll amplitude
Automatically trained and elevated	North reference & elevation angle reference (or signal strength reference in azimuth & elevation	Conical	2 x reference accuracy + other errors	2 x reference accuracy + other errors
Automatically elevated, untrained	Vertical reference only	"Donut" fan	360°	2 x reference accuracy plus other errors
Automatically elevated, Manually trained in azimuth	Vertical reference only	Fan	100° to 180° depending on manual operation availability	2 x reference accuracy plus other errors

reference). One axis stabilization of this type does not eliminate the azimuth errors shown in Figure 2-18 because the system lacks any horizontal or elevation reference. The azimuth error could be tolerated by increasing the horizontal beamwidth of the antenna so that the ship does not "roll" the satellite out of the beam. To determine how wide the beam must be to compensate for Σ, the azimuth error caused by rolling, it is necessary to determine Φ_{rm}, the maximum amplitude of roll. Then, as described in Paragraph 2.4.2, Σ will exceed 90° when ϕ, the satellite elevation angle (measured from horizontal) is

$$\phi \geqq 90° - \Phi_{rm}$$

This marks the point where a satellite on relative bearing of 90° (270°) will cross the zenith of the ships deck plane reference. Since an antenna having a horizon beamwidth of $(\Theta_{hp})_o$ at the horizon may be considered to have a relative beamwidth at other elevation angles of

$$(\Theta_{hp}) = \frac{(\Theta_{hp})_o}{\cos \Sigma} \quad (2\text{-}7)$$

where Σ is the angle above the horizon to which the antenna beam is elevated, the beamwidth necessary to compensate for rolling is then

$$(\Theta_{hp})_o = 180° \cos (90° - \Phi_{rm}) \quad (2\text{-}8)$$

The horizontal beamwidth meeting this criteria will have an elevated beamwidth of 180° at the elevation angle at which the maximum roll causes the azimuth error to approach or exceed 90°. Thus, a fan beam antenna with horizontal beamwidth of 90° and a vertical beamwidth of 60° and elevated to $\phi = 60°$, will tolerate the ±30° rolls on large ships with its vertical beamwidth and will cover 180° of azimuth at elevations of $\phi \geqq 60°$.

The other types of antennas listed in Table 2-5 are discussed in later sections of this report.

Table 2-6 summarizes beamwidth requirements for large and small ships for the types of limited stabilization discussed in this section. From Equation (2-1),

$$|G| = \frac{30,000}{\Theta_h \ \Theta_v} \quad ; \quad |G| < 10 \text{ dB} \qquad (2\text{-}1)$$

the maximum peak gain that can be obtained for an antenna stabilized by a gyrocompass only and satisfying the criteria for beamwidth great enough to accommodate the expected roll on a large ship without major loss of gain, is

$$|G| = \frac{30,000}{90°\text{x}60°} = 5.6, \text{ or } 7.4 \text{ dB}$$

Of course, the operational gain at the roll excursions would be 3 dB less, i.e., 4.4 dB. On a small ship, the gain would be less.

The beamwidths given in Table 2-6 do not account for a vessels average forward speed, or course and speed made over a period of time, which can not be compensated by a gyrocompass reference. The new fast container ships and some special purpose vessels have sustained speeds approaching 30 knots. At 30 knots, a vessel will travel 1330 kilometers in a day. The greatest total change of angle in 3 dimensions is when the ship crosses the subsatellite point on the equator.

Since synchronous satellite altitude is about 35,900 kilometers above the earth's surface, the total angle change is

$$\beta = \frac{1,330}{35,900} = 0.0372 \text{ radians or } 2.13°$$

for the worst case conditions described above. This angle could he all elevation change or made up of components of azimuth change and eleva-

TABLE 2-6

LIMITED MOTION ANTENNA COVERAGE REQUIREMENTS

			Large Ship (30° Roll)		Small Ship (45° Roll)	
Type Antenna	Stabilization Reference	Antenna Beam	Horizontal Beamwidth	Vertical* Beamwidth	Horizontal Beamwidth	Vertical* Beamwidth
Fixed	None	Hemispherical	360°	-25° to +90°	360°	-40° to + 90°
Automatically trained in azimuth, manually elevated	Gyrocompass, North reference only	Wide, Elliptical Cone	90°	60°	130°	90°
Automatically trained in azimuth, manually elevated	Maximum Received Signal Strength	Fan	20°	60°	20°	90°

* Vertical Beamwidth angle is based on the assumption that the antenna will be elevated to angle such that the satellite is in the center of the beam when vessel is upright (0° roll or 0° instantaneous heel)

tion change. To be entirely azimuth change, the satellite would have to be on the ships horizon, in which case the distance to the satellite would be greater by approximately the radius of the earth (6360 km or 3433 n mi.). In this case, the maximum possible azimuth change in 24 hours would be about 1.8°.

It may be concluded, therefore, that an extra 2 to 2.5 degrees in both horizontal and vertical beamwidth will be sufficient to permit manual re-adjustment once each day of an antenna slaved to internal ship references (e.g., compass or inclinometer). This will further limit the gain achievable with a fan beam concept based on azimuth-only slaving. Analogous conclusions may be drawn for concepts based on fan-beams in azimuth which are directed in elevation only. That is, non-symmetrical beams are desirable only if beam-pointing requirements are reduced to a single axis. Since all ships have gyrocompasses, but most do not have elevation or vertical references, it would be most appealing to require only azimuth pointing. However, as shown above, the dynamics of ships are such that such a scheme would result in a severe limitation in operational antenna gain achievable (or undesirable service outages).

Azimuth fan-beams, with only elevation axis directivity, could be configured to permit more gain, however, at the expense of requiring more frequent manual adjustment (re-calibration) of the azimuth direction which must be done to accommodate ships turning as well as its lineal motion (speed). Frequent manual calibration is an undesirable requirement from an overall system reliability standpoint. In addition, the installation of an adequate elevation reference and slave mechanism would be the burden of the antenna subsystem supplier, in contrast to schemes using only the azimuth reference. The various aspects of slaving antenna beams to shipboard references are discussed in Section 3.4.

SECTION 3

TRACKING CONCEPTS - PARAMETRIC CONSIDERATIONS

3.1 BEAM POINTING AND/OR SELECTING

As indicated in Section 2.1, the crux of the shipboard antenna subsystem design problem is the beam pointing mechanism. This is in contrast to other satellite communication terminal applications where the sheer size of the antenna presents an economic imposition, and accordingly aperture efficiency becomes a critical parameter. For the subject MARSAT application, where gains from 3 dB to 18 dB might be considered, aperture efficiency is at best only a secondary consideration, since the antenna size requirements are quite modest at L-Band.

3.1.1 Relative Antenna Size

The relationship between antenna gain and effective aperture A_e is

$$G = \frac{4\pi}{\lambda^2} A_e \qquad (3\text{-}1)$$

where λ is the wavelength of the operating frequency. Since the gain problem of concern here is that for the satellite-to-ship link in the band 1535 MHz to 1542.5 MHz, the wavelength of concern is

$$\lambda = \frac{c}{f} = 19.5 \text{ cm} \quad \text{(or 7.67 inches = 0.64 ft.)} \qquad (3\text{-}2)$$

For a peak gain of 18 dB, the effective aperture area required is 0.19 m^2 (2.05 $ft.^2$). For a peak gain of 12 dB, the area A_e required is one quarter of this, i.e., 0.05 m^2 (0.5 $ft.^2$). The ratio between effective area and actual area is the aperture efficiency, which when various other factors are included, is referred to as antenna efficiency η. In terms of actual area A, then the (peak) antenna gain can be defined as

$$G = \frac{4\pi}{\lambda^2} A\eta \qquad (3\text{-}3)$$

This is illustrated in Figure 3-1 over a range of efficiencies. While many of the antenna radiators discussed in this report do not have circular apertures, the diameter of an equivalent circular aperture can be used as a measure of the linear dimensional change associated with different gains and efficiencies. This is illustrated in Figure 3-2 for the practical range of antenna efficiencies 50% to 70%. These curves demonstrate two points of significance. First, antenna efficiency is not critically important to terminal complexity. The difference between 50% and 70% efficiency would imply a difference in linear dimension of about 5 cm (i.e., 2 inches) for the gains of concern. Secondly, gain itself is not critical in terms of imposing significant dimensions on the antenna, especially for values of gain under 15 dB or so. The imposition on terminal design is then primarily that due to the need for some form of beam pointing mechanism, as was noted in Section 2.1.

3.1.2 Alternative Beam Pointing Techniques

Beam pointing is defined herein very generally to include the selection of different fixed beams as well as controlling a motor-driven pedestal or a network of phase-shifters. The complexity required of the pointing mechanism is certainly a direct function of beamwidth and thus gain, but only in a general sense. Surely, for an antenna gain required of only 3 dB, a simple switch between two beams such as port and starboard, or

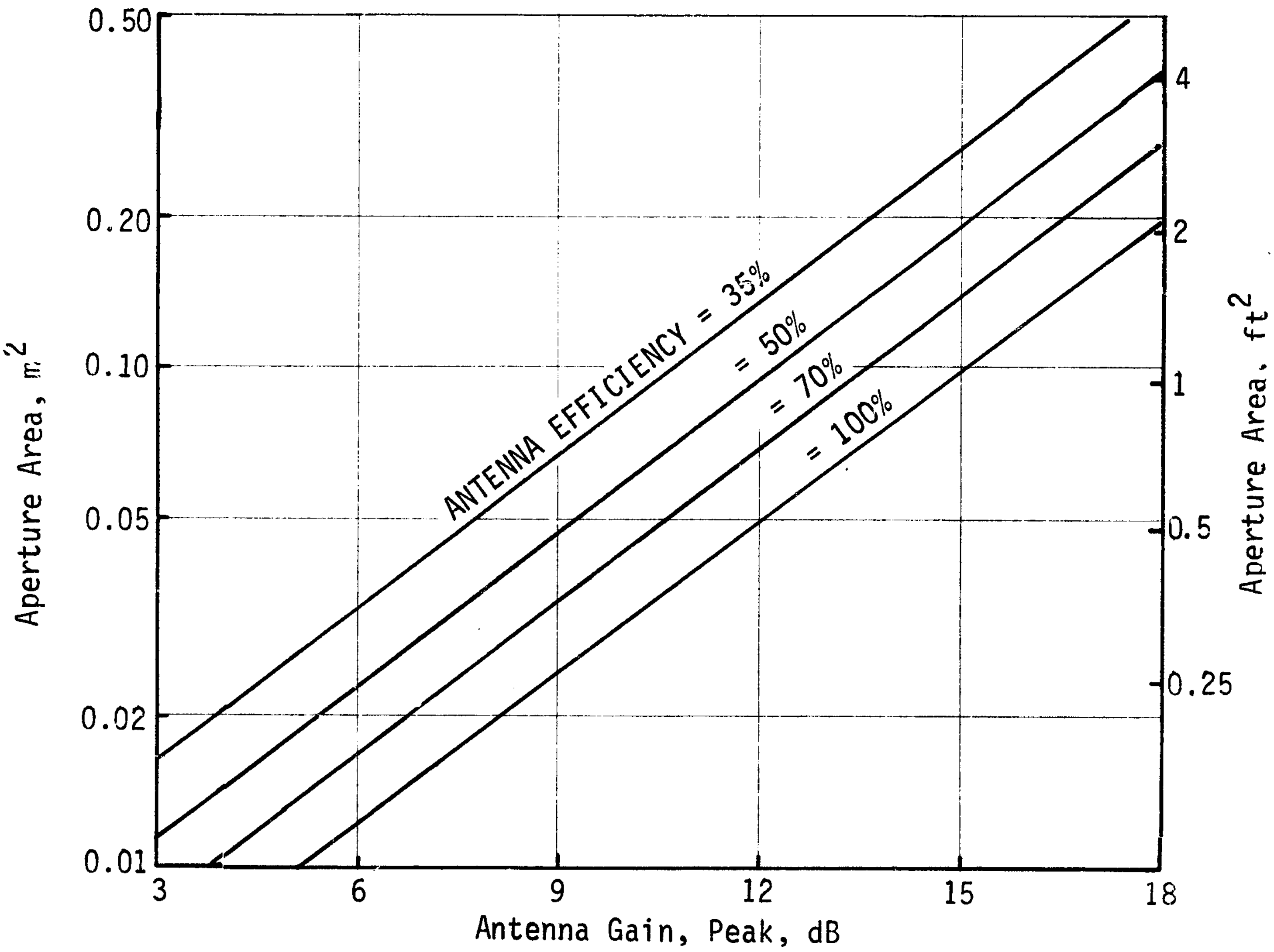

Figure 3-1 RELATIVE ANTENNA SIZE VS GAIN

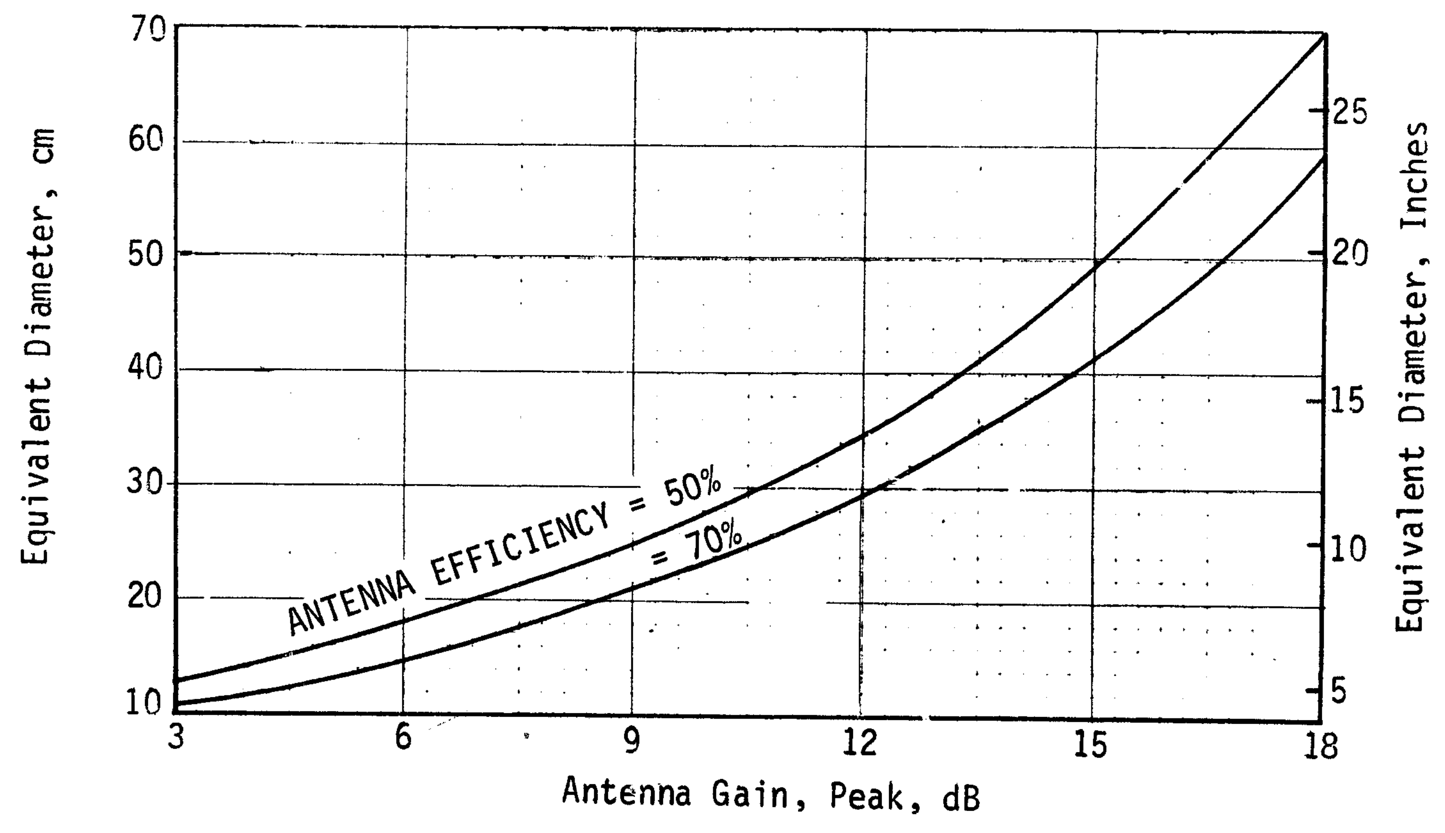

Figure 3-2 EQUIVALENT DIAMETER OF CIRCULAR APERTURE ANTENNA VS GAIN

fore and aft, could readily be performed by the radio operator. Higher values of gain, e.g., 6-7 dB, and thus narrower-beams, would require more frequent switching, to the point where an operator may not be able to effectively accommodate much dynamics in ships motion. At this point, the only options available are those of slaving the beam controller to shipboard inertial references or to a receiver which is locked to a satellite signal.

Slaving to shipboard inertial references is practical, although the frequency with which re-calibration and alignment must be performed increases with decreasing beamwidth, i.e., increasing gain. Gyrocompasses exist on virtually all large ships, and remote repeaters can be used to control antenna beam azimuth angle with an accuracy generally sufficient for the beamwidths being considered ($\geq 20^0$). Elevation references do not exist on most ships, so that use of such a concept would require that the cost of the additional reference be a part of the MARSAT terminal. The cost of elevation references depends on the accuracy vs dynamics performance and generally are not insignificant. Another disadvantage of slaving is that operator calibration is required on a routine periodic basis. While this may be trivial, a totally automatic system would of course be more desirable.

Automatic tracking of the received satellite signal direction is the most common form of beam pointing. Generally, a separate receiver is required, as well as an antenna design which accommodates the developing of a pattern null at the boresight direction. Manual alignment, i.e., critical acquisition, is required only when a ship leaves port or crosses over from one satellite zone to another. The required pattern null may be formed by rotating the antenna beam (sequential lobing) or by forming separate adjacent beams (simultaneous lobing) such that a difference pattern is developed which has a null (zero voltage) in the satellite signal direction.

B

Clearly, both the slaving approach or the automatic tracking approach are applicable to beam controllers which are motor driven gears (mechanical pointing) or digital phase shifters (electronic scanning), or selector switches.

Thus, there are essentially three basic forms of beam pointing control "sensors":

1) Manual tracking
2) Slaving to shipboard inertial references
3) Automatic satellite angle tracking

Also, there are three basic forms of beam controllers, which may be categorized as:

A) Mechanical drive
B) Electronic steering
C) Beam switching

This section generally addresses the pointing control "sensors". A fourth concept is included which is a hybrid of 1) and 3) above, in that it is a simple mechanization of manual tracking which is automatic, but does not require the formation of a separate pattern null.

3.2 EFFECT OF DIRECTIVITY ERROR

Since a primary performance parameter used in the trade-off between beam pointing approaches is accuracy, it is of value here to review the effects of pointing error on terminal performance. The major effect, of course, is to reduce the operational value of antenna gain. The link must be sized for operational rather than peak gain. The amount of effective gain loss due to pointing error is calculated as follows.

The main beam of an antenna pattern may be represented by a power function of gaussian distribution:

$$P(\theta) = V^2 e^{-p^2 \theta^2} \tag{3-3}$$

where V is a voltage (amplitude) factor, "p" is an antenna geometry constant, referred to as the beam shaping factor and θ is the angle off the boresight direction. Equation 3-3 is a fairly precise characterization of pencil (high gain) beams. For certain low gain antennas such as helices, the assumption of a gaussian shape may be somewhat more approximate. However, for purposes of this section, it is considered sufficient. The power loss L_t in dB due to a tracking or pointing error $\Delta\theta$ is given by

$$L_t \text{ (dB)} = 10 \log \frac{P(\Delta\theta)}{V^2} = -10\, p^2 (\Delta\theta)^2 \log e \tag{3-4}$$

At the half-power point $\theta_{hp}/2$ (where θ_{hp} is the two-sided antenna beamwidth) the loss is obviously 3 dB. Thus,

$$p^2 = \frac{3}{10 (\theta_{hp}/2)^2 \log e} \tag{3-5}$$

In terms of half-power beamwidth then, the loss is then

$$L_t \text{ (dB)} = 12 \left(\frac{\Delta\theta}{\theta_{hp}}\right)^2 \tag{3-6}$$

This simple relationship is plotted in Figure 3-3. In typical aperture area limited antennas, allowable tracking error is usually specified to be quite low (typically one-tenth of the half-power beamwidth) so as to keep operational losses low (0.1 dB). For the MARSAT application, it is not nearly so critical in that additional aperture area could be used -

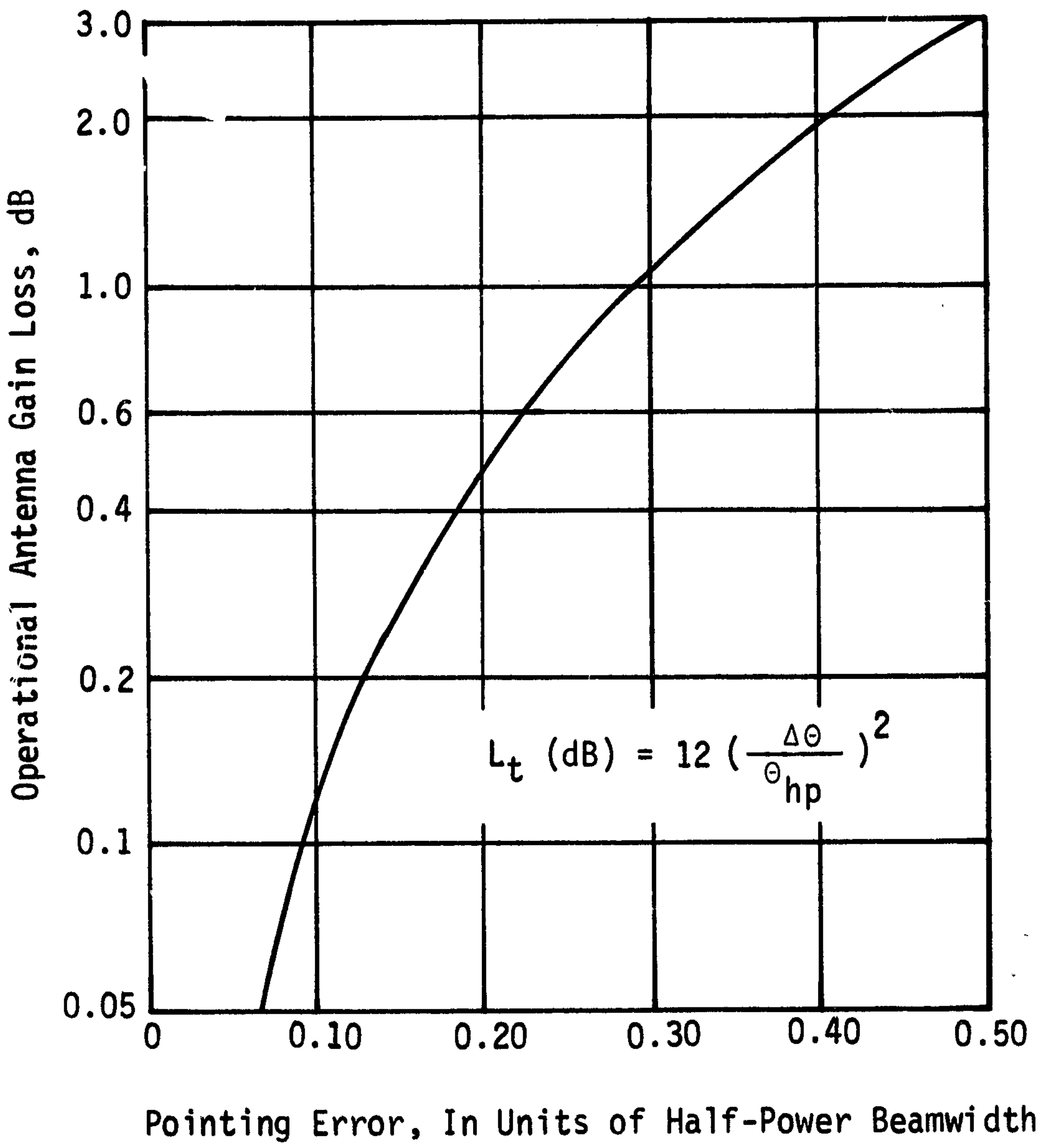

Figure 3-3 EFFECT OF ANTENNA POINTING ERROR ON MARSAT LINK PERFORMANCE

as noted in Section 3.1 - without much imposition in terminal complexity. However, it would only be of value if allowance for large pointing error permits a significant reduction in the cost (or complexity) of the beam pointing system.

Figure 3-3 also illustrates a point with regard to the use of switched-beam antennas. If the individual radiators were to be placed such that beams overlap at the half-power points, each radiator would have to be designed for a peak gain 3 dB higher than that needed for link operation. Overlapping at the 1 dB or 2 dB points on the beam (0.3 or 0.4 times beamwidth respectively) would obviously require more antenna radiators. This trade-off may be extended to show that the number of switchable beams required to achieve a minimum gain of G over the hemisphere is on the order of 2 $|G|$, i.e., twice the absolute (numerical) value of gain.

The effect of pointing error can also be used to indicate potential limitations in gain for antenna subsystem approaches which do not employ continuous pointing control mechanisms to follow the instantaneous dynamics of the ship. That is, one simple antenna approach would be to utilize a beam broad enough to accommodate most (e.g., 99%) roll-angles, and generally employ manual tracking, e.g., once or twice a day, to re-align the beam towards the satellite ("Roll angle" is used loosely here to signify any angular change due to vessel motion and may be the vector composite of motion about all ship axes.) The larger the roll angle potential, the larger is the effect on gain, as indicated in Figure 3-3. For a given roll angle, the higher-gain antennas experience more operational degradation in gain than do lower gain antennas. The question may be asked whether a lower gain antenna can have a higher net gain at a given roll angle than a higher gain antenna. The answer to this question is yes, as shown in Figure 3-4. For example, at an angular offset

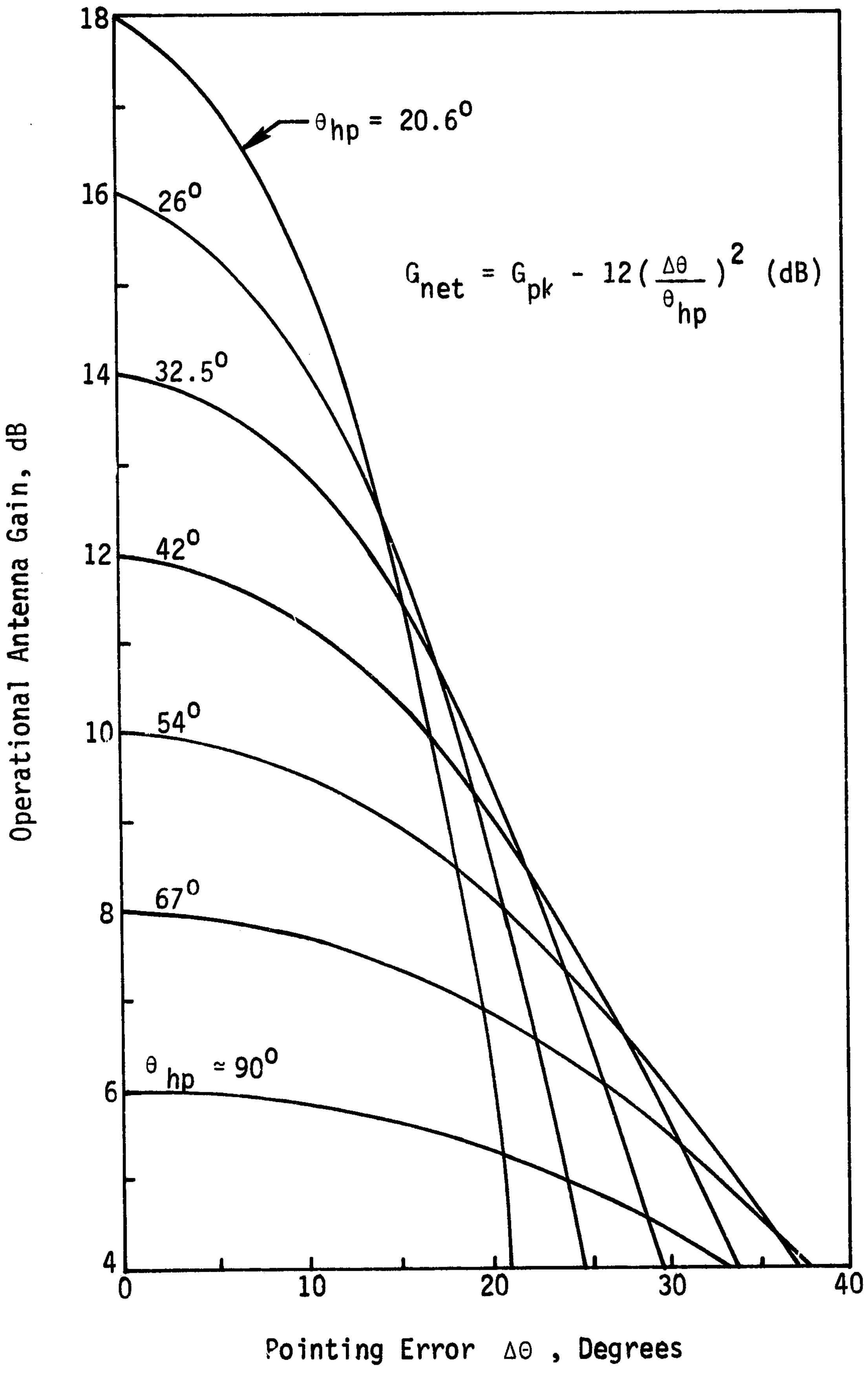

Figure 3-4 NET OPERATIONAL ANTENNA GAIN VS POINTING ERROR FOR VARIOUS ANTENNAS

(pointing error) of 30 degrees, the highest net gain is given by a 10 dB peak gain antenna. It is followed by the 12 dB, 8 dB and 6 dB peak gain antennas, respectively. The following table gives maximum net gain achievable for various pointing offsets, and the peak gain of the antenna which produces it.

TABLE 3-1

MAXIMUM NET OPERATIONAL GAIN VS
POINTING ERROR FOR "FIXED" ANTENNAS
(Symmetrical Beams)

Pointing Error	Maximum Net Gain, dB	Peak Antenna Gain, dB
20^{o}	9.5	14
25^{o}	7.5	12
30^{o}	6.2	10
35^{o}	4.6	10
40^{o}	3.5	8

Figure 3-4 and Table 3-1 apply only to antenna radiators which have symmetrical beam shapes. It is not meant to imply that antenna beams should be selected for the absolute worst-case ship rolls. However, the implications of limited gains for the "nominally-fixed" antenna approach do exist.

Another obvious facet of Figures 3-3 and 3-4 is the effect of beamwidth on potential navigation services which could employ two or more satellites for range-range types of position fixing. The use of the higher gain user antennas, if necessary for link performance, will dictate the need for two independent antenna subsystems on board the ship; or constrain the angular separation of the satellites.

3.3 MANUAL POINTING

As long as an antenna is pointable or selectable, manual pointing is always possible. It may be on the basis of an a priori known direction of the satellite, or on the basis of maximizing received signal level.

For the simple antenna requirement of 2-3 dB gain or so, the least complex approach is obviously the use of two antennas, where selection of port and starboard, or fore and aft beams is enabled by switching and performed by the radio operator. Use of fore and aft beams would probably be preferrable in terms of accommodating the large roll component of ships motion, wherein the operator need switch beams relatively infrequently. Switching would be made on the basis of ships direction relative to the azimuth of the know subsatellite point on the earth. This form of pointing is clearly restricted to antennas with very wide beams, corresponding to gains of 6 dB or less.

The more typical approach is manual tracking, wherein an operator directs (or selects) the beam so as to maximize received signal level. In contrast to seeking the null of a difference pattern (such as provided with a monopulse feed and comparator assembly), conventional manual tracking seeks the peak of the antenna beam. When an operator has visual access to a receive signal level indicator (for example, an S-meter or a receiver AGC voltage monitor), he can, by steering the antenna (for example) until the receive signal level is maximized, generally maintain the antenna beam on the satellite. The radio operator, when manually tracking a satellite, must first acquire the signal and obtain a receiver lock. After ensuring that the lock is firm and that the AGC level has settled, he moves the antenna in one axis and monitors the AGC level. If the signal level increases, the operator continues to move the antenna in the same direction. If it decreases, he reverses the direction and continues to move the antenna until the signal level is maximized.

The process is repeated on the second axis and the antenna is stopped after both axes are optimized. When the AGC level decreases by a given amount, the search is performed again. After the second adjustment, the operator knows which way to move first, thereby making the next search shorter in duration and smaller in amplitude. After two or three moves, the operator can determine the rate, and rather than waiting for a decrease in signal, he anticipates the motion and moves the antenna along the path of the satellite. After this amount of control has been achieved, a major search has to be initiated only when the rate changes significantly or when the operator overcorrects.

Essentially all satellite communication terminals, for all applications, which have steerable beams have the provision inherently for manual tracking, in that some form of receiver output level is usually displayed, and manual control of beam position is always provided for purposes of back-up and/or initial acquisition. The same will be true for the MARSAT terminal if either the slaved pointing or signal autotracking is implemented.

A functional block diagram of the MARSAT terminal system is shown in Figure 3-5. It will probably have at least three receivers, i.e., for access control, voice and data (telex, facsimile). It is anticipated that a special non-working channel will be continously transmitted from each satellite for purposes of system access control, and perhaps other functions such as navigation. The RF carrier in the control channel should serve as a beacon, so that all user terminals can remain locked to the satellite, at least in frequency (AFC) and receiver gain (AGC). If the antenna subsystem includes automatic or manual tracking, it is this signal which should be tracked. While there may be more power in other

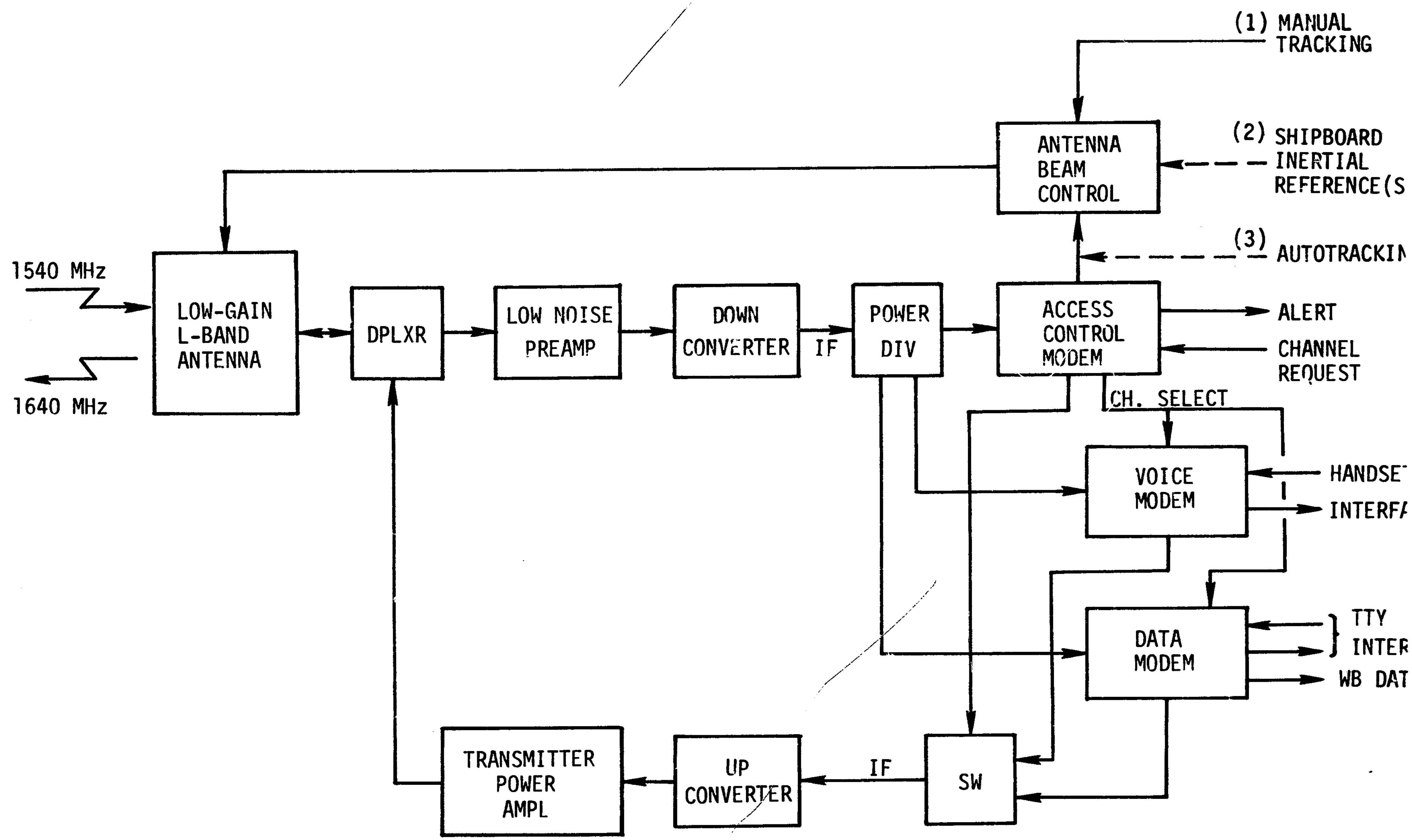

Figure 3-5 MARSAT USER TERMINAL - FUNCTIONAL BLOCK DIAGRAM

(working) channels, their signal bandwidth would be greater, and channel occupancy may vary in center frequency. The control channel signal carrier frequency should be fixed, except for a small amount of system oscillator drift and even less doppler shift. The control signal demodulator will surely be phase-locked, where receiver bandwidth can be relatively narrow, and IF and AGC filter signal-to-noise ratios consequently high.

The disadvantages of manual tracking as a principal mode of pointing, in addition to the obvious fact that it places a burden on the radio operator, include a requirement for slightly higher signal levels and practical unsuitability as a means of compensating for ships motion in rough seas. The higher signal levels are required to discern differences in AGC levels while the antenna beam is considerably offset from the satellite signal direction. In the absence of high angular dynamics, an operator can generally keep the antenna beam pointed at the satellite to an accuracy corresponding to at least twice the AGC level stability. As the latter can typically be as high as 0.5 dB, it can be estimated that the pointing accuracy would be sufficient to maintain the beam or boresight to within the 1 dB points, i.e., ±30% of the half-power beamwidth.

Since the angular rates of motions to be expected with user ships can be quite high (Section 2.4), it must be concluded that manual tracking is unsuitable for the higher antenna gains, i.e., those antennas which have beamwidths which do not accommodate a high percentage of the roll, pitch and yaw amplitudes. The provision for manual tracking should, however, be included in any MARSAT terminal which has a steerable antenna. The basic function would be required for periodic alignment in schemes using slaved pointing, and for initial acquisition and general backup in any system concept.

3.4 ANTENNA POINTING BY SLAVING TO SHIPBOARD DIRECTIONAL REFERENCES

One potentially attractive technique for directing a user ship's satcom antenna beam toward the satellite is by slaving its beam control mechanism to directional references installed on the ship, with periodic alignments made on the basis of received signal level. Depending upon the beam pattern of the antenna, references in one, two and three dimensions may be required, as noted in Section 2.5.

The location of the satellite relative to the ship is defined (for purposes of pointing the antenna's axis of maximum gain at it) by its azimuth and its elevation. Azimuth is defined herein conventionally as the angle in the horizon plane as measured clockwise from the North (the direction of the earth's North pole). Elevation is conventionally measured up from the horizon plane to the zenith on the azimuth (bearing) of the elevated object.

There is only one candidate to be considered for an internal north reference on a ship. This is the compass. Two types of compasses are available on most ships. They are gyrocompasses and magnetic compasses. The gyrocompass is the more suitable of the two because it furnishes north reference directly without correction for magnetic variation, caused by non-coincidence of the earth's magnetic and true north poles, and with less deviation, potentially caused by the magnetic fields of the ship itself. In addition, the usual shipboard gyrocompass has a repeater system already installed which furnishes north reference directional information to various locations on the ship. The gyrocompass is typically used on the ships likely to be early subscribers to a maritime services satellite system.

Candidate sensors for the automatic continuous output of elevation angle reference on a merchant ship are not as obvious. The only sensor universally used is the seaman's eye looking at the horizon. This is clearly not automatic nor is it continuous. The accelerations and velocities of a ship's motion in a seaway make an elevation angle reference difficult. On land, surveyors have used gravity in the form of a plumb bob or pendulum for a vertical reference in the determination of elevation angle since the beginning of cartography. The dynamic motion of a ship makes the use of a pendulum, or inclinometer, more difficult. However, it is one of the directional reference sensor candidates. The other candidate is a vertical gyro. The vertical gyro is a complex device which makes it susceptible to failure. It is also expensive, and not typically installed on ships postulated to be satcom users. Further, its accuracy is not generally warranted for the subject application.

The accuracies of angular references, the implications of installing such devices in ships, the costs, the advantages and disadvantages, and the effects of ships motion on operation and accuracy are discussed in this section.

3.4.1 Horizontal Angular Reference

As noted above, the shipboard gyrocompass is the only practical candidate for azimuthal angular reference on-board ship. Not only is it almost universally installed but also it has readily available in its repeater system the means of furnishing directional references to remote locations.

Gyrocompasses used on merchant ships possess reasonable accuracy. Most of them are provided with manual latitude and speed correction. The navigator sets in his latitude and speed as changes occur by means of controls on the master compass which is usually located in the chart room area just aft of the wheelhouse. With corrections entered, average quality gyrocompasses (such as Sperry's Mark XIV) will have errors of less than 1^{o}. Speed changes can cause errors of the order of $1/2^{o}$ if they are drastic (e.g., from 15 knots to 0 knots). Improper latitude correction or operation at high latitudes (e.g., North of 60^{o} N.) will cause additional errors. Experience has shown that the root mean square error will be less than 2^{o}, 99% of the time. This is adequate for pointing of antennas whose horizontal beamwidth is 5^{o} or more, providing the servo system which trains the antenna does not introduce excessive errors. If greater accuracy is required there is a means for providing it within reason. The navigator routinely checks the accuracy of his gyrocompass by comparing an observed azimuth of the sun, a star or a range (two terrestrial objects, the line of bearing through them being known and observable) with a computed azimuth of the same objects. The difference is the gyro error which can be applied to the azimuth setting of the MARSAT antenna. These procedures can reduce the azimuth reference error to less then 1°, 99% of the time.

Modern gyrocompasses use synchros for their repeater systems. With this approach, a synchro generator, driven by the master compass follow-up system, in turn drives the synchro motors in the compass repeaters. An amplifier is used in the line between the master and its repeaters to provide power to drive a limited number of repeaters. It is possible to use the north reference synchro voltages as one input to a synchro differential generator whose other input may be manual and whose output controls a servo motor which trains the antenna to the desired horizontal azimuth, as shown in Figure 3-6. The manual input is the true bearing (azimuth) of the satellite.

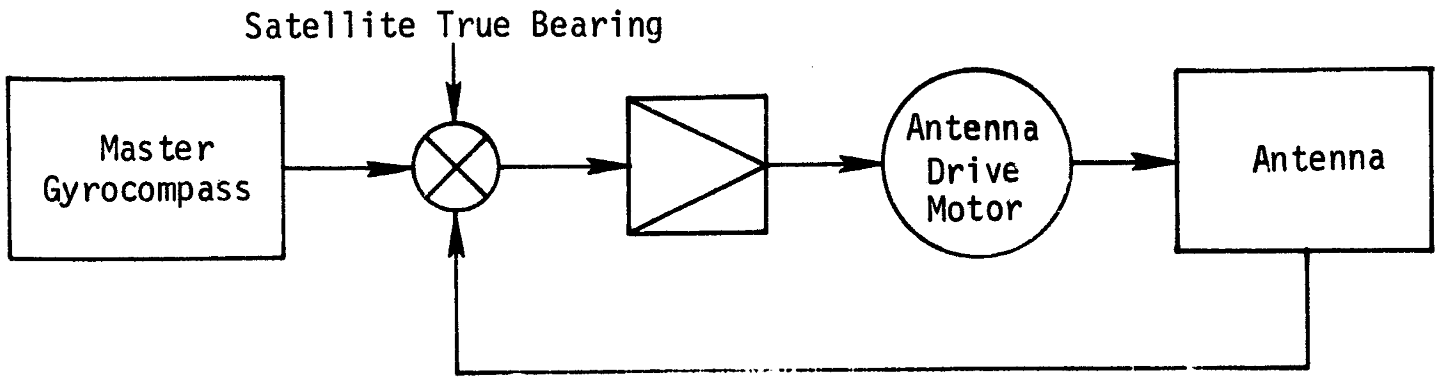

Figure 3-6 GYROCOMPASS-SLAVED ANTENNA CONCEPT

If the ship's gyrocompass has a "step-by-step" repeater system, it would either be necessary to convert to a synchronous system with an additional unit, or control the servo motor directly from a step motor. The step-by-step repeater system is an older system still found on many ships. In this system, as the ship turns the master compass follow-up generates pulses which "step" the repeaters in small increments of a degree (1/6 deg.). Every time a repeater is shut off and re-energized, it is necessary to re-align it with the master. This procedure is cumbersome and may result in additional errors due to faulty alignment.

If a ship is proceding in relatively calm water so that the amplitudes of roll and pitch are small (10° or less) and the antenna is elevated to the computed elevation angle of the satellite, a fan beam antenna of 10° to 20° horizontal beamwidth can be trained by azimuth information from the gyrocompass without any elevation angle reference and the satellite will remain within the solid angle illuminated by the antenna. As shown in Sections 2.4 and 2.5, when roll angle amplitudes approach the maximum expected, even for large ships, an antenna controlled in azimuth only by north reference can be trained to the wrong relative bearing at the peak of the roll. The amplitude of this azimuth error is a function of the relative azimuth and elevation of the satellite. As shown in Section 2.5, an antenna stabilized in azimuth only (north reference) and unstabilized in elevation (no horizon or vertical reference) needs a horizontal beamwidth of approximately 90° to continuously illuminate a satellite at any elevation angle from a ship rolling with an amplitude of ±30°.

3.4.2 Vertical or Elevation Angle References

There are two prime candidates for a self-contained or internal elevation angle reference. They are:

- Vertical gyroscopes
- Pendulums

The elevation angle reference is the reference direction in a vertical plane from which the actual angle of the satellite above the horizontal plane is determined. It may be in the horizontal plane or it may be at some angle to the horizontal plane. Except for visual references, the most common reference direction is the vertical. The force of gravity is commonly used to determine the vertical. A gimbal mounted pendulum may be used as a reference. On a platform moving at constant velocity it furnishes a reference precise enough for most navigational and survey purposes. When subject to acceleration such as results from rolling and pitching of a ship in a seaway, errors occur in the vertical as indicated by the pendulum. This is illustrated in Figure 3-7 in which the antenna at the masthead is gimbal mounted with its center of gravity below the axis of the gimbal bearings. It is not drawn to scale. The antenna and its gimbals are exaggerated in size.

3.4.2.1 Simple Pendulum Type Vertical References

A pendulum or gravity type vertical reference is the simplest type vertical reference used on ships. Many ships are fitted with inclinometers which are instruments used to measure the angle of heel or instantaneous angle of roll and most of these instruments are a simple, undamped pendulum of about 10 cm length. Theoretically, the pendulum, with the force of gravity exerted on it, will remain vertical while the ship rolls under it. The pendulum has a pointer at its tip which, as the ship rolls, indicates the angle of roll on the graduated scale. An inclinometer is one candidate for a stabilization sensor.

Several possibilities exist for pendulum stabilization, where the pendulum may rotate in either one or two axes:

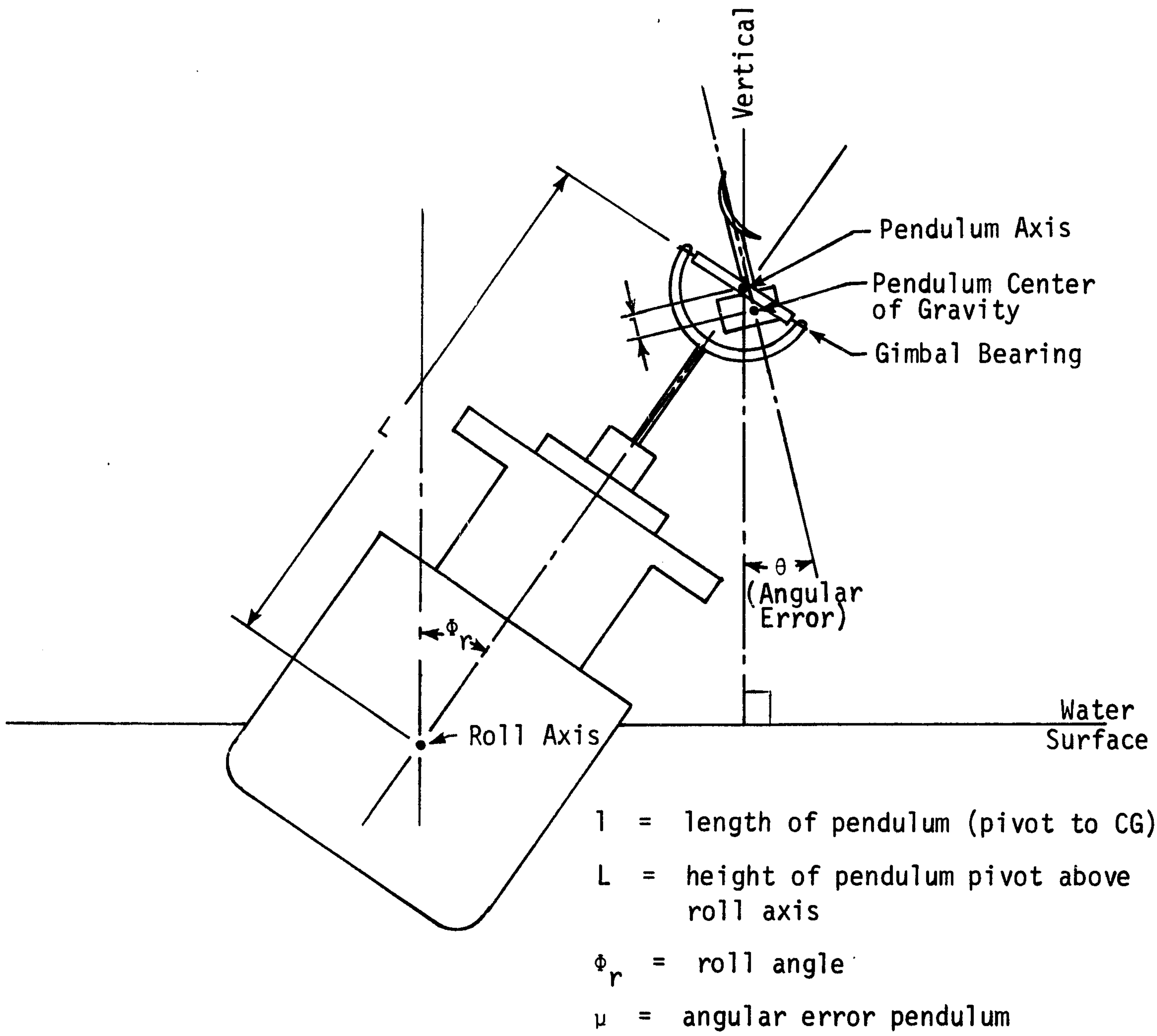

Figure 3-7 GIMBALLED MASTHEAD MOUNTED SATCOM ANTENNA WITH ANTENNA SIZE EXAGGERATED TO SHOW ANGLE AND DISTANCE RELATIONSHIPS ON A ROLLING SHIP

- Antenna gimbal-mounted on one or two axes with center of antenna gravity below gimbal axis
- Pendulum located at a distance from antenna and used to sense roll or roll and pitch for control of the antenna.

Whether the pendulum should be free to rotate on one or two axes depends on the magnitude of the roll and pitch angles expected. The previously noted CCIR paper[30] gives a maximum expected pitch angle for a vessel 50 meters long as 15°. The probability of a pitch angle this large however is 1 in 10^8. Since vessels almost never encounter synchronous pitching and the maximum expected effective wave slope is 9°,[27][29] the maximum expected pitch angle is taken as 9° rather than 15°. This applies only for small vessels, less than 100 feet in length. Large vessels, such as 1000 foot long supertankers will span more than a wave length under most conditions and will not pitch through amplitudes of more than 2°. The antenna to be stabilized should have a beamwidth which is adequate to keep the satellite illuminated even with the greatest pitch amplitude anticipated. The conclusion may be drawn therefore that stabilization around only the roll axis is required and a simple one-axis inclinometer is the prime candidate.

Mounting an antenna in gimbals as shown in Figure 3-7 is an attractive idea because it makes a self-contained unit. If this unit is at the top of the mast, as shown, it could be subjected to accelerations of more than twice that of gravity due to the rolling. Since gravity is the force determining the vertical, errors will result. Friction in the bearings or added damping will affect the angular error (μ in Figure 3-7) of this pendulum. The angular error (μ) can be determined

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(30) CCIR, Doc. M/241-E. Op. Cit.

(27) Neuman and Pierson. Op. Cit.

(29) Comstock, J. P. Op. Cit.

by summing the moments or torques acting on this pendulum system. This process may be characterized by the following differential equation:

$$m\, l^2\, \ddot{\mu} + c\dot{\mu} + mgl\mu = ml\, L\, \ddot{\Phi}_r + c\, \dot{\Phi}_r \tag{3-7}$$

in which

m = pendulum mass

g = acceleration due to gravity

c = coefficient of damping (bearing resistance)

There are three solutions for this equation which are pertinent:

- Undamped case, $c = 0$
- Critically damped case, pendulum damped to eliminate pendulum oscillation at its natural frequency
- The general case, pendulum oscillates at its natural frequency and the roll frequency, ω_1

Using LaPlace Transforms to determine transient and steady state response[32] and transforming variables yields the following equations:

The general case:

$$\frac{\mu(t)}{\Phi_r} = \frac{h}{g}\left[\frac{(L\omega_1{}^2 h)^2 + (2rg)^2}{(h^2-1)+(2rh)^2}\right]^{\frac{1}{2}} \sin(\omega_1 t + \psi) - \frac{h}{g\sqrt{1-r^2}}\left[\frac{(2r^2+L\omega_1{}^2)^2+(2rg^2\sqrt{1-r^2})^2}{(1-h^2[1-2r^2])^2+(2h^2r\sqrt{1-r^2})^2}\right]^{\frac{1}{2}}$$

$$\times \left[e^{-hr\omega_1 t} \quad \sin\,(h\omega_1\sqrt{1-r^2}\,t - \psi')\right] \tag{3-8}$$

where

$$\psi = -\left(\tan^{-1}\frac{2rg}{Lh\omega_1{}^2} + \tan^{-1}\frac{2rh}{h^2-1}\right) \tag{3-9}$$

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(32) Gardner and Barnes. Transients in Linear Systems. John Wiley and Sons, 1942.

$$\psi' = -\left(\tan^{-1} \frac{2rg\sqrt{1-r^2}}{2r^2g+L\omega_1^2} - \tan^{-1} \frac{2r\sqrt{1-r^2}}{h^2-1+2r^2}\right) \tag{3-10}$$

The undamped case:

$$\frac{\mu(t)}{\Phi_r} = \frac{Lh^2\omega_1^2}{g(h^2-1)} \left(\sin \omega_1 t - \frac{1}{h} \sin h\omega_1 t\right) \tag{3-11}$$

The critically damped case:

$$\frac{\mu(t)}{\Phi_r} = \frac{\omega_1 h}{g(h^2+1)} \left[L^2\omega_1^2h^2 + \frac{4g^2}{\omega_1^2}\right]^{\frac{1}{2}} \sin(\omega_1 t + \psi_c)$$

$$- \left[(2g + L\omega_1^2)ht - \frac{2}{\omega_1} \left[\frac{g-(g+L\omega_1^2)\, h^2}{h^2+1}\right]\right] e^{-h\omega_1 t} \tag{3-12}$$

$$\psi_c = -\left(\tan^{-1} \frac{2g}{L\omega_1^2 h} + \tan^{-1} \frac{2h}{h^2-1}\right) \tag{3-13}$$

In these equations:

Φ_r = amplitude of roll

$\mu(t)$ = instantaneous vertical error as a function of time (t)

ω_1 = roll frequency in radians/sec; $\omega_1 = 2\pi/\tau_r$, where τ_r is the roll period in seconds

h = β_o/ω_1 where β_o is the natural frequency of the pendulum, $\beta_o = \sqrt{\frac{g}{l}}$ where l is the effective length of the pendulum

g = acceleration of gravity, 32.2 ft/sec.^{2}

L = height of pendulum axis above the ship's axis of roll

r = c/C_c where C_c is the coefficient of damping for the case of critical damping, $C_c = 2\ ml\ \sqrt{gl}$

Examination of these equations show that when L is large, the errors are large regardless of how much damping is applied to the pendulum or how the length and hence the natural frequency of the pendulum is varied. In fact, in the undamped case, the error is proportional to L and is zero when L is zero. Figure 3-8 is an example of the normalized instantaneous errors for a critically damped pendulous antenna mounted at the mast head of a large ship. It indicates that when the pendulum frequency is large compared to the roll frequency ($h > 1$) the transient response dies out in less than 1/4 of the roll period and the steady state response is the largest error. This occurs at a time lagging the maximum roll angle by about a second or 1/12 the roll period.

In the general case, illustrated in Figure 3-9, the error function describes a more complex curve. This curve represents the normalized errors of a short pendulum (l = 39.4 cm) at the center of roll with minimal damping ($r = 0.1$). Here the maximum error is an order of magnitude smaller than when L = 43 meters, and the maximum error occurs about 1/2 the pendulum period after initiation. This is the first point at which the transient oscillation instantaneous amplitude peak adds to the steady state instantaneous amplitude. Since a large h causes the transient to die out sooner, the indication is that h should be as large as possible, i.e., that the pendulum length be small.

Figure 3-10 illustrates the limitations of the critically damped pendulum. The only quantities which can be varied are L, the height above the roll axis and h, the ratio of pendulum to roll frequencies. Variation of the pendulum length within practical limits has very little effect on the magnitude of vertical error. The curve indicates that minimum error occurs when $L = 0$, as would be expected.

Figure 3-11 shows the variation of roll-induced pendulum or inclinometer error when the pendulum is in the vicinity of the roll axis of a ship.

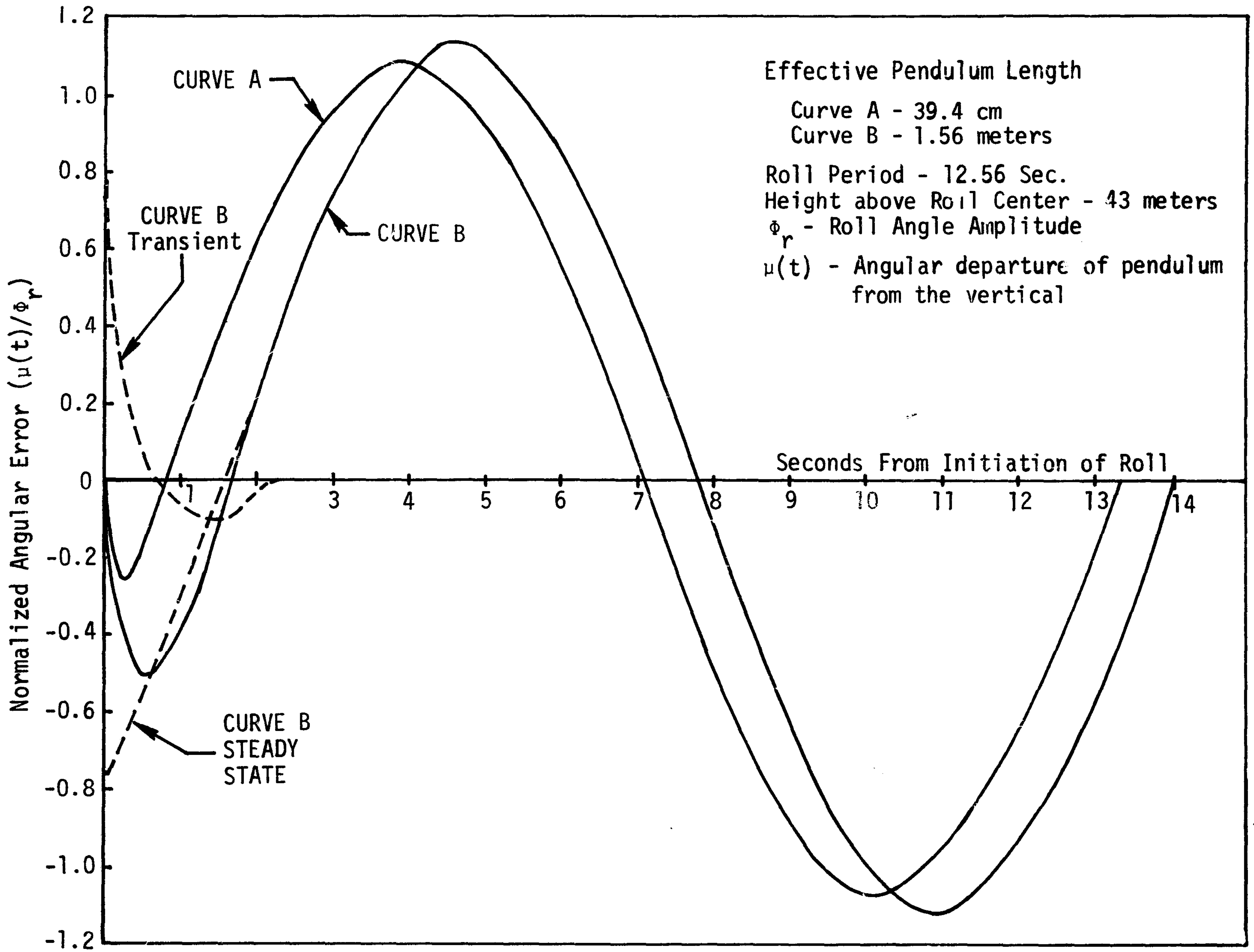

Figure 3-8 GIMBAL MOUNTED ANTENNA, CRITICALLY DAMPED. ANGULAR ERROR OF VERTICAL CAUSED BY ROLL - LARGE SHIP

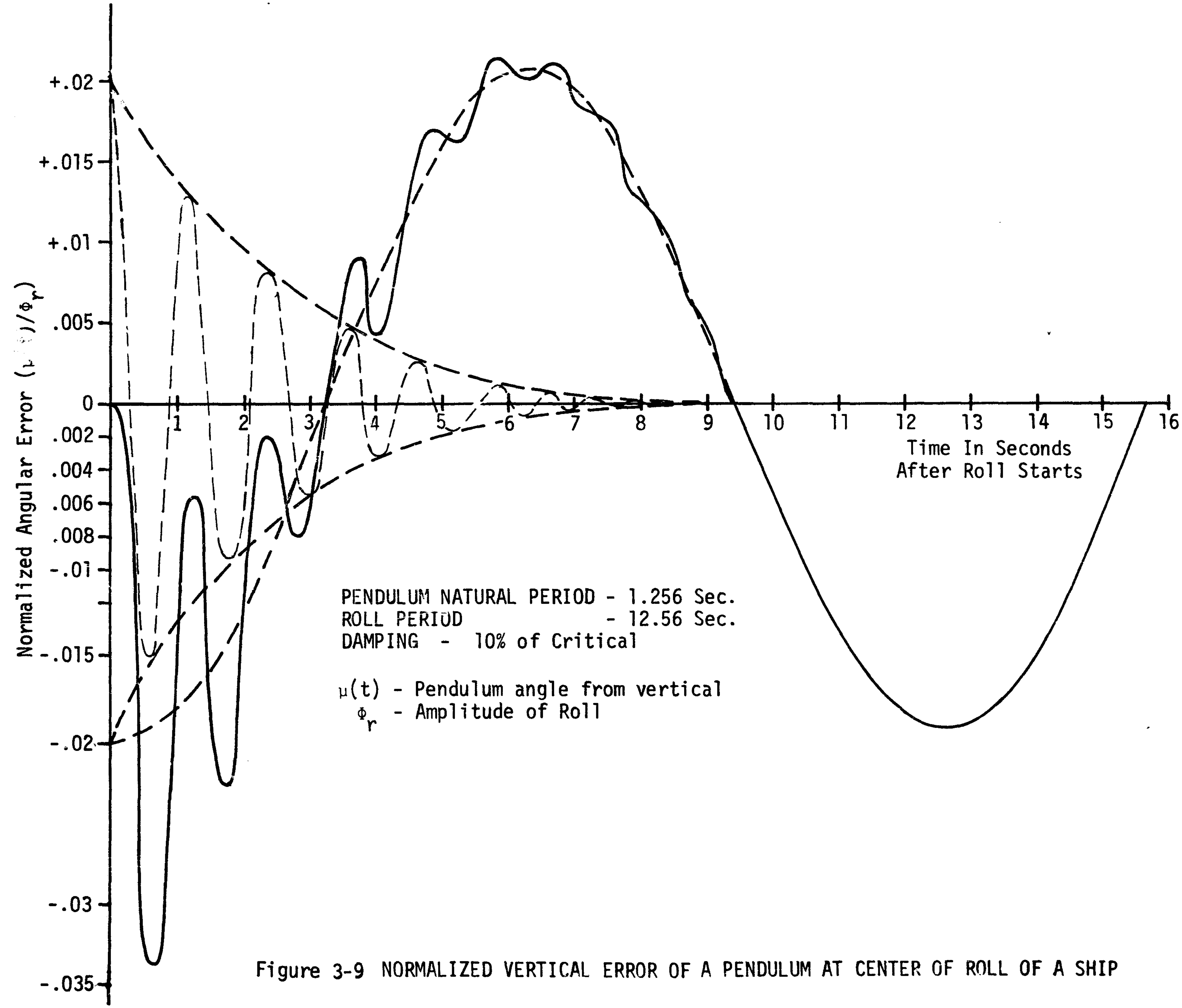

Figure 3-9 NORMALIZED VERTICAL ERROR OF A PENDULUM AT CENTER OF ROLL OF A SHIP

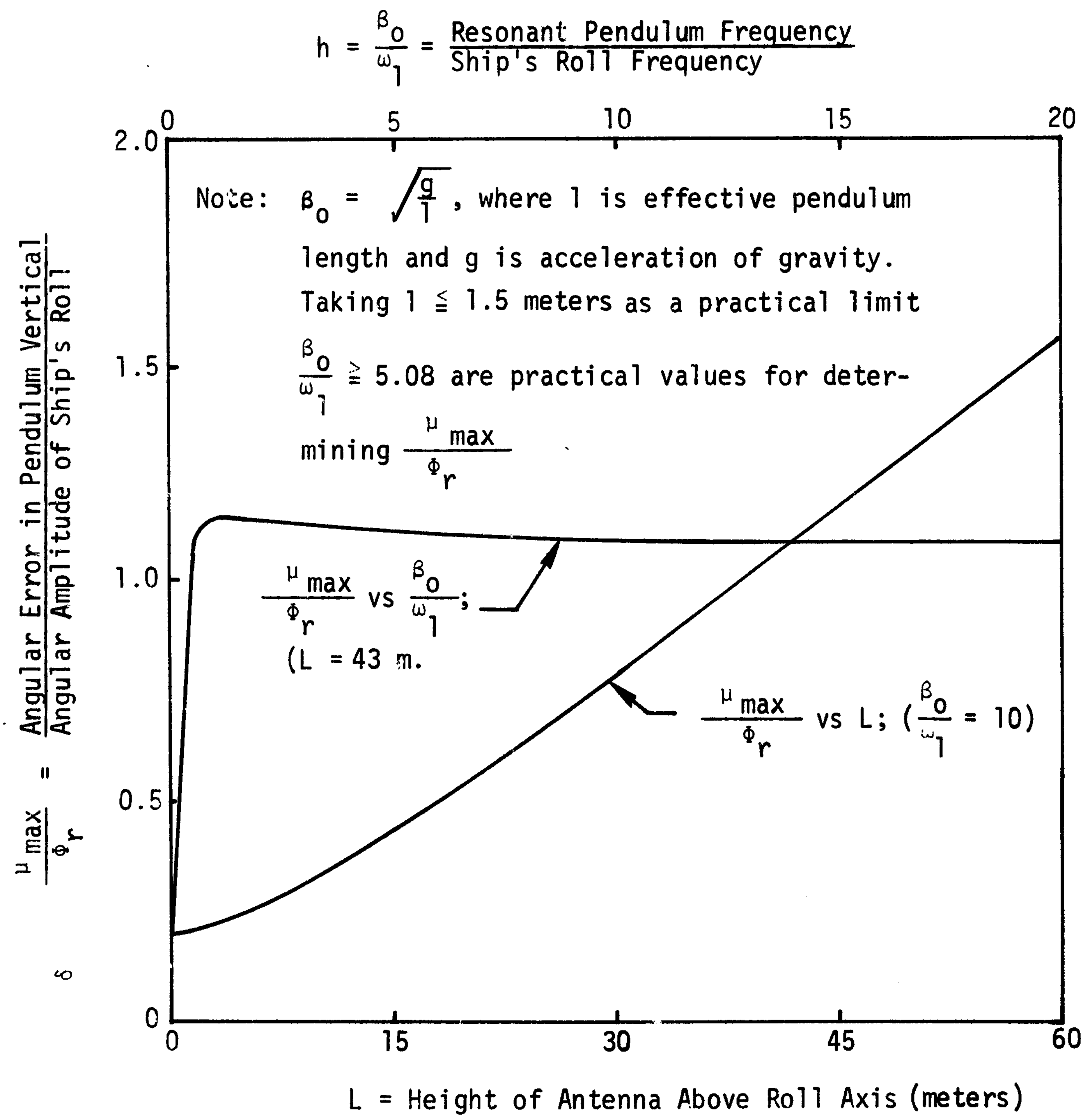

Figure 3-10 MAXIMUM ANGULAR ERROR OF A PENDULUM VERTICAL REFERENCE AS ITS HEIGHT ABOVE THE SHIP'S ROLL AXIS AND ITS RESONANT FREQUENCY ARE VARIED

CRITICALLY DAMPED CASE

ω_1 = 0.5 radians/sec. (Roll Period = 12.56 sec.)

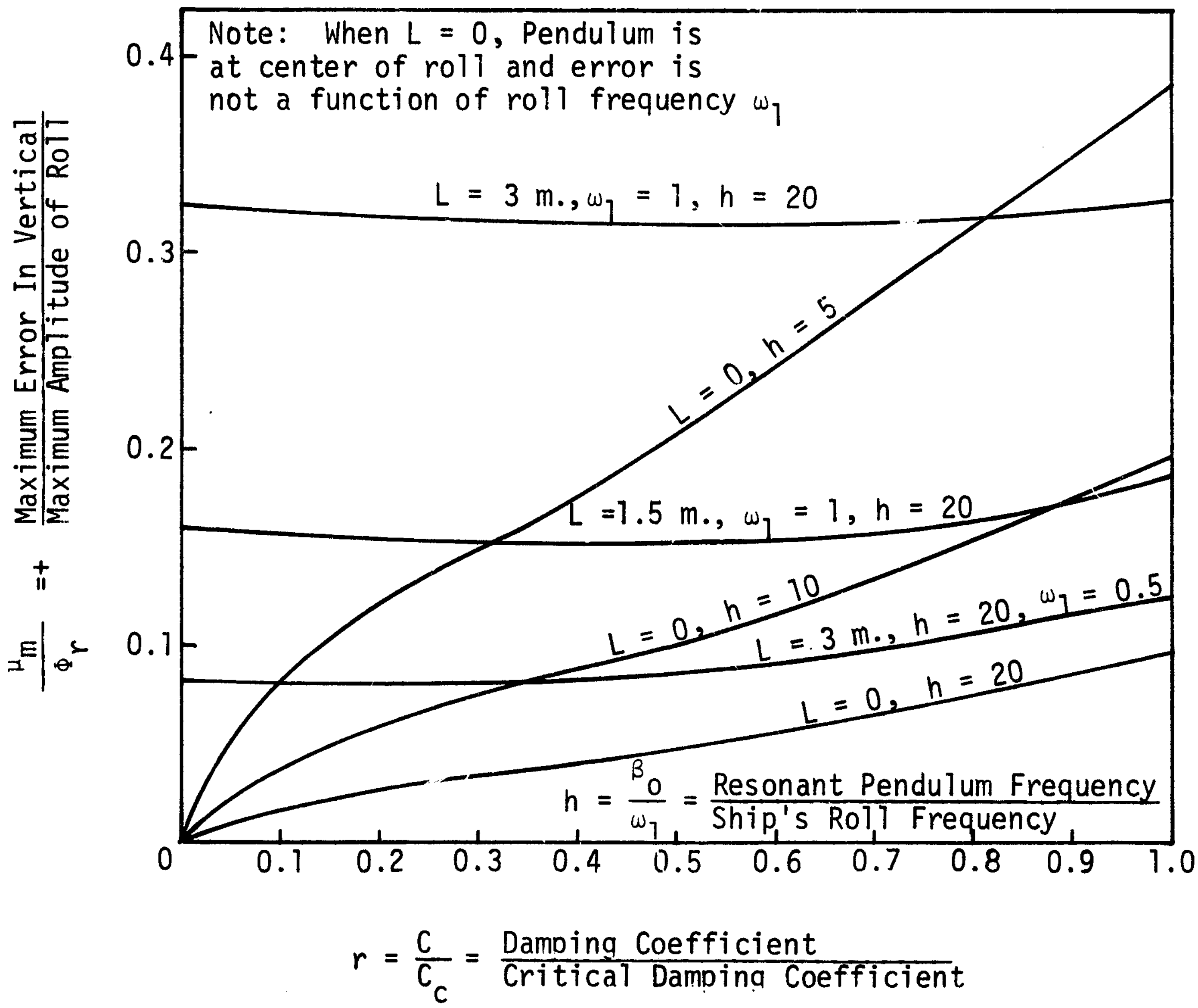

Figure 3-11 NORMALIZED ERROR OF PENDULUM VERTICAL IN THE VICINITY OF A SHIPS ROLL AXIS AS A FUNCTION OF PENDULUM DAMPING

If the inclinometer could be installed on the roll axis with a minimum amount of damping and a length such that its resonant frequency is 20 times as great as the ships natural frequency of rolling, errors of less than 3% can be obtained. This means the pendulum length will be a little less than 1 inch for the "small" ship characteristics and a little less than 4 inches for the "large" ship characteristics, as cited in Paragraph 2.4.2. The three curves in Figure 3-11 for L = 0 cover pendulum lengths from 39.4 cm to 2.4 cm for the small ship, whose resonant roll period is 6.28 seconds, and from 1.56 meters to 9.8 cm for the large ship, whose resonant roll period is 12.56 seconds. The upward hump in each of the L = 0 curves between r = 0 and r = 0.5 is due to the addition of the transient error response and the steady state error response when they are in phase.

Unfortunately, the axis of roll which corresponds approximately to the transverse center of gravity[(29)] does not remain at the same height above the keel in ships which carry cargo. A shift of the order of 6 meters may be expected in a supertanker, for instance. If the center of gravity varies from 3 meters above to 3 meters below the pendulum location, the curve for L = 10, h = 20, ω_1 = 0.5 shows that less than 10% error in vertical may be anticipated. The small vessels with a higher maximum roll frequency incur greater vertical reference errors with a movement of the center of gravity. A movement of 3 meters can cause an error of better than 30%. Fortunately, the shift of the center of gravity on a small ship is much smaller than on the large one. The L = 1.5 meters curve indicates that errors less than 20% may be expected. An alternative which could be used to minimize errors in vertical reference is to install more than one pendulum sensor at different heights above the keel.

The conclusion of this analysis is that a single-axis pendulum or inclinometer to sense roll angle is a practical means of stabilizing an

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(29) Comstock, J.P. Op. Cit. pg. 670.

antenna whose beamwidth is approximately 40° or more. A pendulum integral with the antenna itself is not practical because the antenna will necessarily be in the upper part of the ship. A remote pendulum vertical sensor located near the ship roll axis can be used, however, with errors in the range from 3% to 20%.

Implementation of an inclinometer stabilization system involves a servo system similar to the yaw (azimuth) axis stabilization system described in Section 3.4.1. The simplest system is illustrated in Figure 3-12.

In this system the angle of roll is transmitted from the inclinometer by means of a synchro which controls a servo motor on the antenna. The motor positions the antenna to the angle measured from the deck (reference) plane as indicated by the inclinometer.

The inclinometer used for a sensor should meet several constraints to make it as accurate as possible. It should:

- Be located as close to the center of gravity as possible. (Use of two or more may be desirable.)
- Have a short effective length, about 2.5 cm for small ships and about 10 cm for large ships
- Have slight damping, 10 to 20% of the critical. Since $C_c = 2mg^{\frac{1}{2}}l^{\frac{3}{2}}$ and the length l is determined by other considerations, a fairly heavy pendulum is indicated so that bearing friction and the loading of the synchro will not cause excessive damping.

3.4.2.2 Vertical Gyros

Vertical gyros have been used in aircraft and military ships for stabilization for a number of years. The primary use in military ships is as an accurate reference to remove errors in gun laying or missile firing.

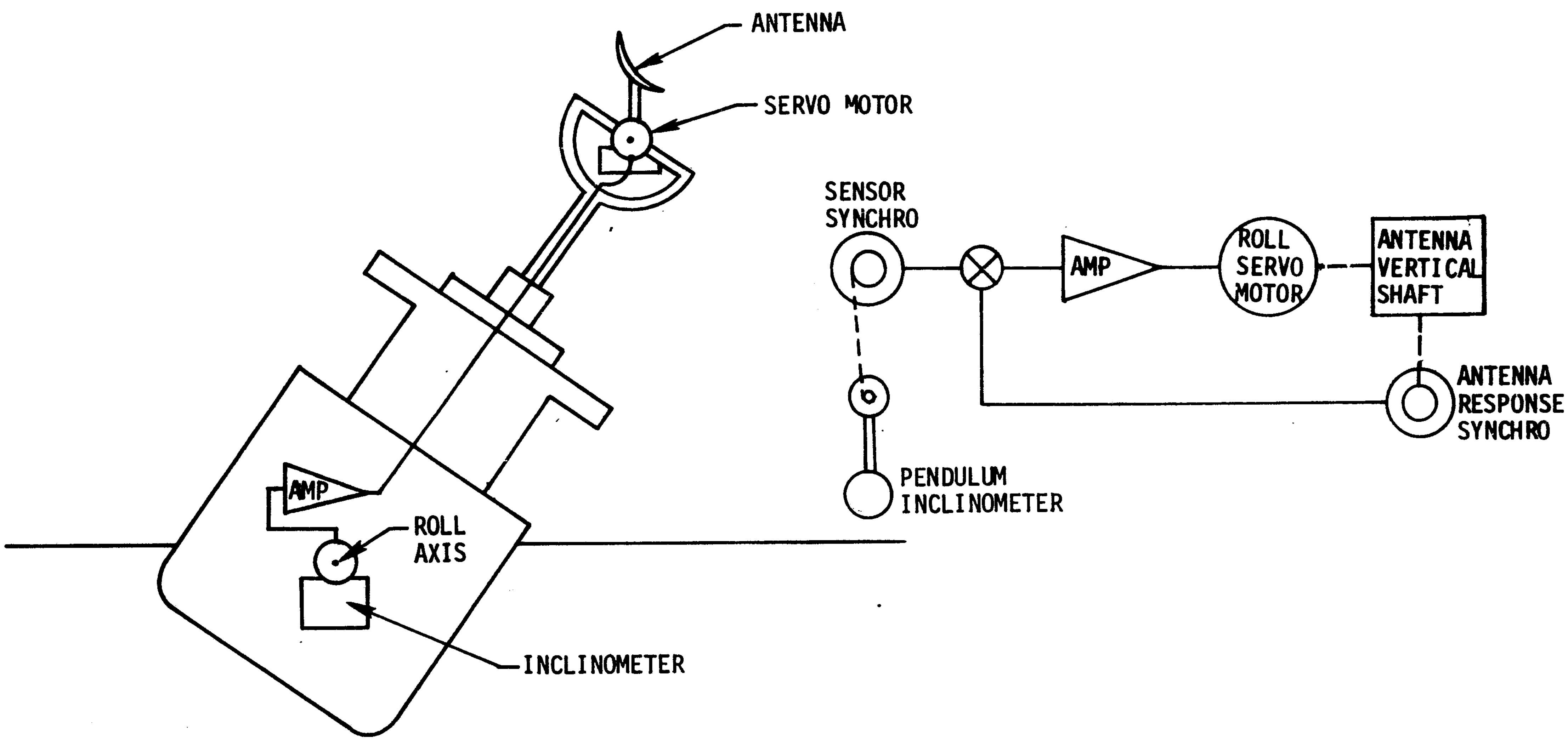

Figure 3-12 ANTENNA STABILIZATION SCHEME USING A REMOTE INCLINOMETER

As such, these units are designed for intermittent use.

A vertical gyro is a two degree-of-freedom device with gimbal bearings perpendicular to each other. Rotation about these axes determine roll and pitch as measured from the local vertical. The rotor spin axis of the gyroscope is maintained parallel to the local gravity vertical by precessing the gyro with control signals generated by a gravity sensing device, e.g., a viscously damped pendulum. The gyro furnishes short term stability and the pendulum furnishes long term stability (for several minutes to several hours during which the gyro would be offset from the vertical by the rotation of the earth).

Typical errors in vertical reference in a shipboard vertical gyro even 15 meters above the roll axis with 20^{o} roll amplitude is $\pm 0.2^{o}$. This represents a much greater accuracy of vertical reference than required for an antenna with 40° beamwidth. These gyros are typically designed for 1000 to 2500 hours minimum life (0.11 to 0.28 years of continuous operation). This is not sufficient for continuous operation at sea on a commercial vessel.

While vertical gyros are a commercially available device widely used in aircraft, it is concluded that those presently available are not a practical device for merchant ship satcom antenna stabilization; not only because of their complexity but also because of their short design life and [illegible]h periodic maintenance requirements.

3.4.2.3 Comparative Costs of Vertical References

Vertical gyros presently available for shipboard use cost about $4800, not including installation. The simple inclinometer roll stabilization system described in 3.4.2.1 is not yet commercially available. The comparable parts of this system consist of the inclinometer, a synchro and an amplifier. It is estimated that these items can be manufactured and assembled into a unit in quantity to sell for less than $500 per unit. If either were used to stabilize an antenna in roll as shown schematically in Figure 3-12, then a roll servo motor, shaft support, bearings and a gimbal ring will be necessary at the antenna in addition to the normal train and elevation shaft and bearing arrangement required to rotate and elevate the antenna. Produced in quantity, it is estimated that these items would add about $300 to the cost of the antenna. Cost of installation of additional cable will be about $7/foot and may be as much as $560 for either installation. Foundations and other installation costs for the basic references will be higher for the vertical gyro than the inclinometer. These are estimated at $1000 and $200 respectively. The comparative costs of the two stabilization systems are thus:

	Inclinometer	Vertical Gyro
Basic Reference Sensor Cost	$500	$4800
Extra Antenna Bearings & Gimbal Ring	300	300
Installation - additional cabling	560	560
Installation - of vertical sensor	200	1000
Totals	$1560	$6660

While the vertical gyro stable element is much more accurate and precise as a sensor for a satcom antenna, the inclinometer stable element is projected to be entirely adequate in accuracy, more reliable, simpler and, when installed, less costly by a factor of four.

3.4.3 Conclusions on Control of Elevation and Azimuth of An Antenna Stabilized by Slaving

As shown in this section, an unstabilized antenna in effect has its axis of highest gain referenced to the deck plane and the bow of the ship on which it is installed. Use of a gyrocompass for azimuth reference stabilizes the antenna for the ships yawing or turning motion, but does nothing for roll and pitch. Rolling introduces azimuth errors which are large when the satellite elevation is high. Pitching on ships introduces relatively small errors compared to the anticipated beamwidths of a satcom antenna. Hence, stabilization in pitch can be dispensed with. Stabilization around the ship's roll axis is beneficial and azimuth errors are reduced to tolerable limits by the use of a vertical reference having not more than a 20% error. (This corresponds to a 6° error for a large ship and a 9° error for a small ship for maximum roll amplitudes of 30° and 45° respectively.)

When an antenna is stabilized in azimuth and roll, it will remain with the axis of its beam directed at the bearing and elevation (referred to North and the horizon) to which it is set (within error limits) as the ship moves. However, it is still necessary to determine the azimuth and elevation of the relative (desired) geostationary satellite position and direct the antenna beam toward it. This calibration procedure is required on a periodic bases (e.g., daily) and can be done manually by training and elevating until the desired output from the satcom receiver is maximized, following the initial acquisition which can be achieved by manually training and elevating to the computed azimuth and elevation of the satellite.

Since the greatest expected angular change per day for a geostationary satellite is less than 2.5°, a satcom antenna stabilized in azimuth and roll to ±10° and with a beamwidth of 40°, can be readjusted manually on a daily basis without significant loss of efficiency. Adjusting the antenna for maximum receiver output (peak) would be similar to tuning a transmitter or receiver, a procedure with which radio officers on merchant ships are very familiar. It would require azimuth and elevation motors on the antenna and controls at the satcom terminal location in the radio room. Local training and elevation of the antenna, either mechanically or electrically, as a means of making daily adjustments is not considered feasible for an operational installation because on many ships the antenna will be located at the masthead and on occasion the weather will be too bad to permit a man to climb the mast every day to adjust the antenna.

While pointing directive antenna beams by slaving to shipboard inertial references appears to offer the advantage of simplicity, the technique has certain disadvantages. It is susceptible to relatively large pointing errors, and thus degradations in operational gain, and installation is not always as simple as it first seems. Also, visual and audio alarms should be incorporated in the communications receiver (e.g., in the access control channel demodulator) to indicate when received signal levels fall below a prescribed amount.

Most significant, however, is the need for periodic manual realignment. This is not only inconvenient for the crew; it is potentially unreliable. In a worldwide system with many thousands of remote users, it is important that all subscribers are continuously "on the air". The maritime community, e.g., ship owners and the U.S. Coast Guard, must be confident that it can call any ship in the system, and not be dependent on the attentiveness of crew members. The potential cost effectiveness of the slaving approach, however, may warrant its use for at least certain classes of users.

3.5 DEVELOPING A PATTERN NULL - CONVENTIONAL AUTOTRACKING OF SATELLITE SIGNAL

3.5.1 General Alternatives

From a performance standpoint, in terms of accuracy and of reliably maintaining a value of operational transmit and receive antenna gain, no substitute for automatic angle tracking of the satellite signal exists which is better than second-best. It has generally been considered undesirable for mobile satellite applications because of presumed complexity and resultant cost. This presumption is founded in the vast experience with antennas of much higher gain and various other constraints. In addition, few, if any, autotracking systems have been developed for volume production and commercial use.

Given that the antenna beam has to be pointed or selected, either mechanically or electronically, the presumption of added complexity inherent in autotracking is not necessarily valid. The basic additions required are a special antenna feed and/or null-pattern-forming L-Band circuitry, a tracking receiver, and servo electronics.

The microwave null-forming circuitry is definitely an addition, but with today's technology, it by no means needs to be complex. The tracking receiver function can, for certain autotrack concepts, largely be accommodated in the basic access/control signal receiver and demodulator, with minor additional low-frequency modules required. The servo electronics required, to a large extent, will exist in any case if the beam is to be steered, pointed or selected, whether by slaving or remote manual tracking.

The additional terminal complexity associated with the tracking receiver function is the major factor involved in the pointing "sensor" conceptual comparisons. In this context, the most fundamental trade-off consists in the number of channels that should be used in the tracking receiver. Historically, this amounts to a comparison of monopulse (simultaneous lobing) and conical-scan (sequential lobing).[(22)(33)(34)] Classic monopulse can employ either (signal) amplitude or phase comparison schemes, but in any case generally requires a multi-port feed and multiple receiver channels, usually one each for the sum (used for general communications functions) and difference (e.g., azimuth and elevation) channels. Conical-scan implies mechanical nutation (scanning) of a single-feed element (or reflector) about the boresight axis and requires but a single receiver channel, but typically suffers from amplitude modulation interference (scintillation) and penalizes the communication channel, or operational terminal G/T, by the con-scan cross-over loss (i.e., due to the fact that the single feed element is offset from the boresight axis and thus the focal point). Due primarily to the latter penalty and the undesirability of mechanically rotating feed components, monopulse techniques are far more prevalent in current satellite communications antennas. This is also true in telemetry applications such as NASA's Deep Space Network, the (Apollo) Manned Space Flight Network (USBS), the U.S. Air Force's Space-Ground-Link Subsystems (SGLS), as well as all commercial communication (INTELSAT) terminals.

In the INTELSAT terminal application, conical-scan angle tracking is generally not used, since it is not permitted to amplitude modulate the up-link carriers. This is one very distinct characteristic of conical-scan tracking in which a single antenna (feed) is used to transmit as

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(22) Skolnik, M.I. Op. Cit.

(33) Rhodes, D.R. "Introduction to Monopulse". McGraw-Hill Book Co., Inc. New York, 1959.

(34) Stephenson, J., et al, "Design Criteria for a Large Multipurpose Tracking Antenna". Philco-WDL-TR1368, 20 Jan. 1961.

well as receive. While the AM modulation index that would exist in the transmitted signal is very small, it would nevertheless be undesirable for a satellite transponder which was power-limited and thus requires careful power-balancing of carriers. While this is not strictly the case in the MARSAT (ship-to-shore) application (assuming the transponder is cross-strapped with SHF), the sum of the various disadvantages tend to negate the further consideration of conventional mechanical conical-scan angle tracking in this study.

A hybrid scheme employing a monopulse feed but only a single receiver channel has many desirable features. This scheme is alternately referred to as pseudo-conscan, single-channel monopulse or simply "monoscan". The major merit of the scheme is, as noted, the requirement for only a single receiver channel. As a receiver channel generally consists of a preamplifier, axes cable wraps or rotary joints, down-converters, IF amplifiers, filters, detectors, etc., there is a costs savings with the monoscan technique that can be significant. A second advantage is that there is no requirement for the intra-channel phase alignment associated with conventional monopulse. Also, as the scanning function is implemented in the receiver path following the diplexer, there is no amplitude modulation of the transmit carrier. The disadvantages of the monoscan scheme are that terminal G/T is degraded due to the addition of a difference channel combiner and modulator circuitry prior to the low noise preamplifier, and that its pointing accuracy is less than that of conventional multi-channel monopulse for the same received signal level. Two-channel monopulse, in which the two difference signals are combined for processing in a single receiver channel, is also a credible scheme having some of the advantages and disadvantages of both the above.

These trade-offs are examined in the following paragraphs.

3.5.2 Conventional Monopulse

Monopulse tracking is implemented by using either a multiple-port antenna feed or multiple antennas to permit the forming of sum and difference signal patterns. Adjacent patterns on opposite sides of the antenna's boresight direction are compared (subtracted) in a circuit referred to as a comparator in order to form the difference pattern, producing a sharp null in the boresight direction. Both amplitude- and phase-comparison monopulse techniques are possible. For the subject MARSAT application, only conventional amplitude-comparison monopulse is considered appropriate.

3.5.2.1 Configurational Factors

Figure 3-13 shows a functional block diagram of a monopulse comparator circuit. The center point of the aperture of the four-port feed is the antenna boresight. A, B, C and D represent signal levels from the four feed elements. The monopulse comparator is a passive device designed for low-insertion loss. For the narrow-band 1540 MHz MARSAT application, the comparator circuitry is quite conventional and poses no design problems. It is important that the proper phase relationship be established and RF phase errors be eliminated, so that accuracy will not be affected by relative changes in the phase of IF circuitry.

The feed elements, classically multiple horns in large antennas (and more recently multimode single horns), could be crossed-dipoles over a reflector radiator such as the short-backfire antenna or separate spirals in a cavity-backed spiral or helix. The antenna category, of course, must be amenable to the use of separate ports. If tracking in only one dimension was needed, such as for a fan beam, then only two feed ports would be required.

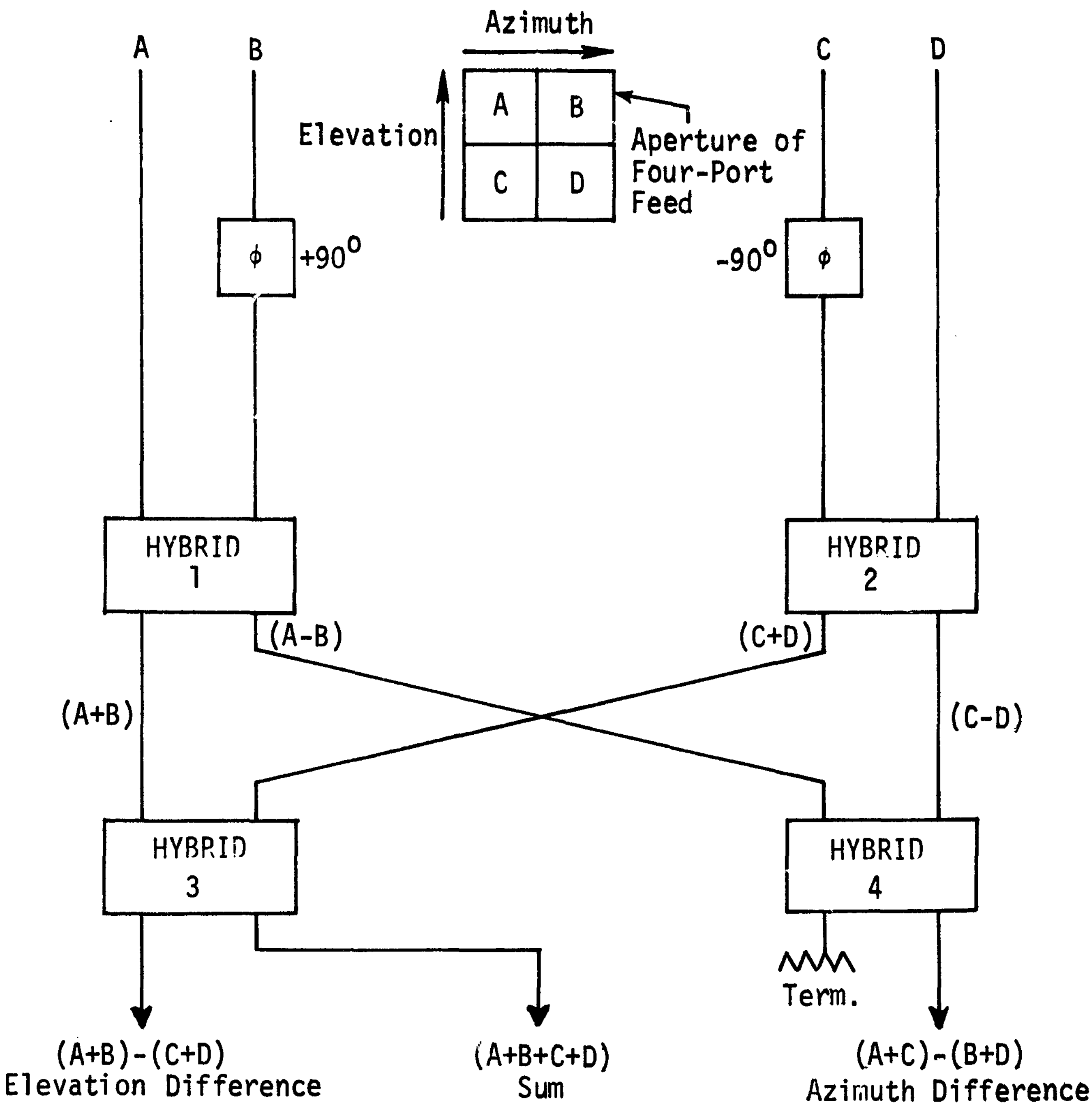

Figure 3-13 MONOPULSE COMPARATOR FUNCTIONAL BLOCK DIAGRAM

The function of the tracking receiver is to down-convert, filter and amplify the low level difference pattern signals, independently, and normalize them to the sum pattern level, producing an error signal output at dc given by

$$\text{dc error signal} = m \frac{E_{\Delta}(\theta)}{E_{\Sigma}(\theta)} \tag{3-14}$$

where m is a receiver (gain) constant, $E_{\Delta}(\theta)$ is difference signal amplitude (either azimuth or elevation) for a given angular error θ and $E_{\Sigma}(\theta)$ is the sum signal amplitude. Figure 3-14 shows a block diagram of a conventional monopulse tracking receiver.

Three channels are typically required, in order to independently process the azimuth, elevation and sum pattern signals. Normalization of the azimuth and elevation difference channel signal levels is accomplished via slaving of the channel gain at IF to the sum channel's AGC circuit. As indicated in Figure 3-14, the basic sum channel would be constituted by the terminal's communication receiver. More specifically, it would be primarily designed for the access control channel demodulation function. The sum channel L-band low-noise preamplifier and down-converter would be of sufficient bandwidth (e.g., 8.5 MHz) to process all communication channels, with channelization achieved at either the 1st or 2nd IF. A triple-conversion receiver is shown because such should be expected to be more compatible with demodulator design for such narrow-band signals. The access control channel demodulator shown is a narrow-band phase-locked receiver, such as would be appropriate for a PSK/PM (subcarrier) control channel modulation.

If this control signal employs direct-carrier PSK modulation, an IF signal squarer would be inserted at the output of the 3rd IF amplifier, the coherent (quadrature) AGC detector could be changed to a non-coherent

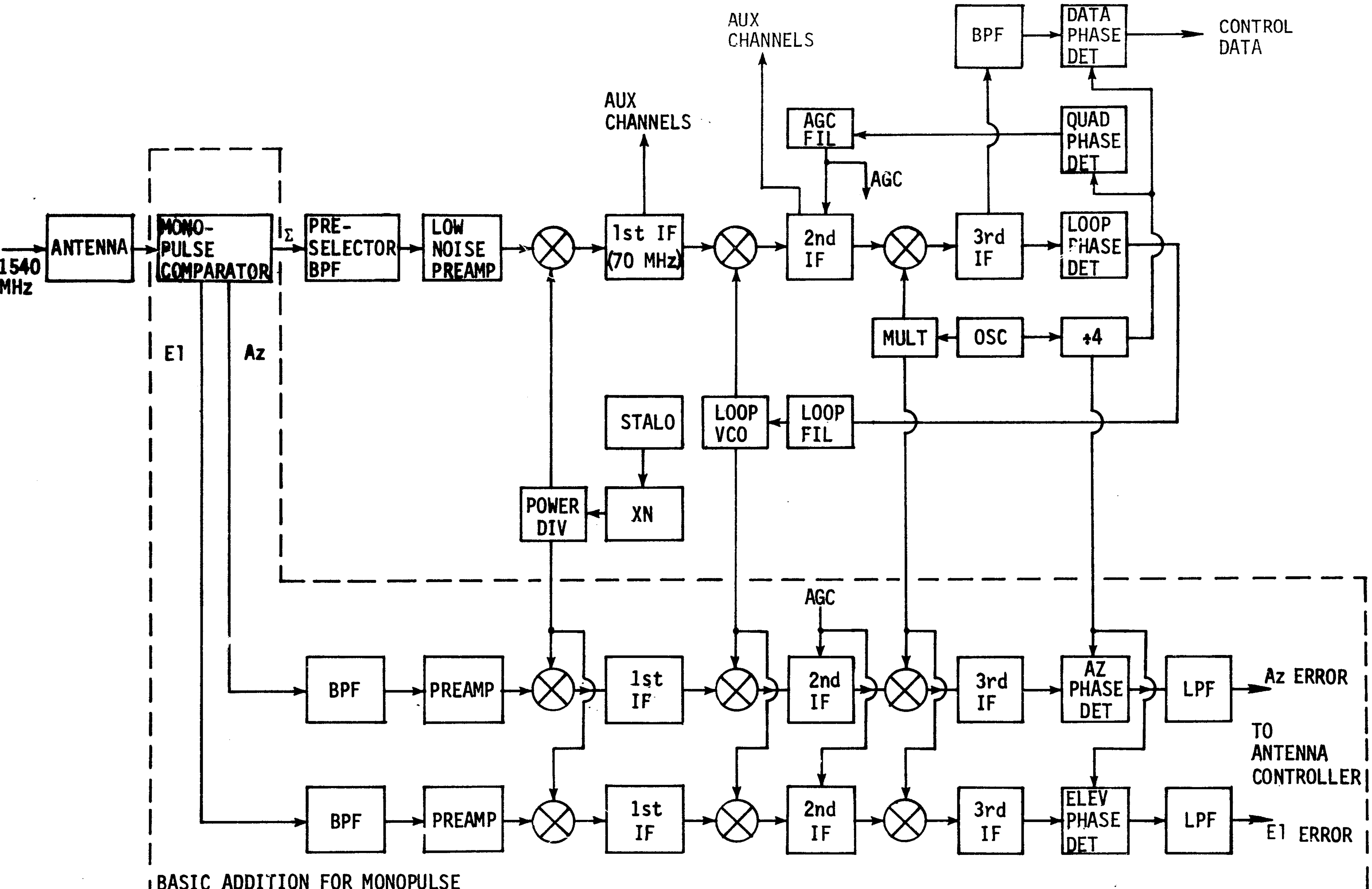

Figure 3-14 THREE-CHANNEL MONOPULSE RECEIVER EXAMPLE

(envelope) detector, and the loop phase-detector reference oscillator would be doubled. Otherwise, the receiver block diagram is applicable to both modulation techniques. (Actually, for direct-carrier PSK, it would be preferrable not to use signal squarers in the difference channel 3rd IF's. This can be alleviated by employing the sum channel's 3rd IF output, after narrow-band filtering but before squaring, as the detector reference in the difference channels instead of the phase-locked VCO. This scheme amounts to conventional cross-correlation tracking of the signal energy.)

The basic point is, however, that two additional receiver channels are required. The amplifiers, mixers and filters, etc., are all duplicates of the sum-channel components. Nevertheless, it is additional circuitry. Assuming the L-band down-converters are antenna mounted, the additional rack-mounted circuitry may typically be a number of cards corresponding to half a standard 18 cm (7 inch) chassis' capacity. The local oscillators, AFC and AGC functions are all supplied by the sum channel.

The additional L-band preamplifiers and down-converter mixers are also an undesirable imposition on terminal equipment. It is not fundamental, however, that monopulse difference channel pre-amplifiers need to have the same low-noise or gain performance as the sum channel. If the closed-loop antenna steering servo is sufficiently narrow-band, a relatively high difference channel noise temperature can be tolerated without significantly affecting tracking accuracy. Some gain may be required however to avoid a requirement for a high degree of isolation in the first L.O. power divider. This isolation must be greater than the sum of the difference in preamplifier gains and the monopulse null depth (relative to the sum channel peak), in order to negate sum channel leakage into the difference channel via the down-converter reference. As the null depth may be 30-40 dB, no more than about 10 dB of difference in preamplifier gains should be used, since more than 40 or 50 dB of L.O. isolation is undesirable.

Since phase alignment between sum and difference channels is also important, in addition to isolation, there may be another motivation to use the same low-noise preamplifiers in all channels. However, cost savings must be considered. The major criterion is tracking accuracy.

3.5.2.2 Tracking Accuracy of Monopulse

Angle tracking errors in autotracking systems are comprised of errors from the following general sources

1) Receiver thermal noise
2) Angular input dynamics
3) Wind gust torques
4) Phase and amplitude unbalance

The most fundamental limitation in angle tracking accuracy is that due to receiver thermal noise, which can have an impact on configuration requirements.

a) Thermal Noise Tracking Error

The thermal noise autotracking error in a conventional monopulse system, where the sum channel signal is used as a (cross-correlation) reference to detect the difference channel tracking error signals, is shown in Appendix A to be given by

$$\sigma_{\theta n}^{2} = \frac{1}{k_{\Delta}^{2}} \left(\frac{1 + \rho_{\Sigma}}{\rho_{\Sigma}^{2}} \right) \frac{B_{\Delta}}{B_{IF}} \; \frac{T_{\Delta}}{T_{\Sigma}} \tag{3-15}$$

where $\sigma_{\theta n}$ = rms tracking error due to thermal noise (deg),

k_{Δ} = Slope of the error channel voltage (normalized by AGC), in volts/volt-deg

ρ_{Σ} = Sum channel carrier-to-noise ratio, C_{Σ}/N_{Σ}, in bandwidth B_{IF}

B_Δ = Two-sided position servo loop noise bandwidth, Hz
B_{IF} = Receiver reference bandwidth (such as IF), Hz
T_Δ = Difference channel system noise temperature, K, and
T_Σ = Sum channel system noise temperature, K.

At high input carrier-to-noise ratios, such as may be assumed when the sum channel signal reference used is a phase-locked VCO, $\rho_\Sigma >> 1$, the following relationship results:

$$\sigma_{\theta n}^{2} = \frac{1}{k_\Delta^2 \rho_\Sigma} \left(\frac{B_\Delta \; T_\Delta}{B_{IF} \; T_\Sigma} \right) = \frac{1}{k_\Delta^2} \left(\frac{K \, T_\Delta \, B_\Delta}{C_\Sigma} \right) \qquad (3\text{-}16)$$

where K, in this case, is Boltzmann's constant.

The normalized slope of the difference channel voltage pattern, at the output of the comparator, is a figure of merit for monopulse feed systems with reference to potential tracking accuracy. It is a characteristic of the feed and antenna geometry only. For a classic four-element monopulse feed, there is an optimum value of normalized error signal slope; that which maximizes the product of sum channel gain at boresight and (absolute) difference channel slope. Appendix B contains a description and derivation of this optimum value, the result of which is

$$k_\Delta = \frac{1.17}{\theta_{hp}} \quad \text{volts/volt-deg.} \qquad (3\text{-}17)$$

It must be re-iterated that this "optimum" value applies primarily to classic four-element feed systems only, where tracking accuracy is of significant value, e.g., in radar.

For the MARSAT user antenna subsystem, tracking accuracy in itself is not important, and as the above corresponds to a feed offset angle which tends to compromise sum channel efficiency, the error channel slope can be allowed to be degraded somewhat in favor of directly maximizing sum channel gain. Even with the corresponding relative improvements in (sum channel) antenna feed efficiency achieved by the space communications antenna industry, however, it is somewhat remarkable that very creditable monopulse error channel sensitivities have still been achieved. Depending on specific antenna types, a typical value of monopulse error channel slopes might be expected to be in the range

$$k_{\Delta} = \frac{m}{\theta_{hp}} \quad , \quad 0.5 < m < 1.0 \tag{3-18}$$

Equation (3-15) may then be written as

$$\left(\frac{\sigma_{\theta n}}{\theta_{hp}} \right)^2 = \frac{1}{m^2} \left(\frac{1 + \rho_{\Sigma}}{\rho_{\Sigma}^{\ 2}} \right) \frac{B_{\Delta} \ T_{\Delta}}{B_{IF} \ T_{\Sigma}} \tag{3-19}$$

The use of Equation (3-6) then permits the above to be expressed directly in terms of effective antenna gain loss due to the (one-sigma) tracking error; i.e., for monopulse,

$$L_t \text{ (dB)} = \frac{12}{m^2} \left(\frac{1 + \rho_{\Sigma}}{\rho_{\Sigma}^{\ 2}} \right) \frac{B_{\Delta} \ T_{\Delta}}{B_{IF} \ T_{\Sigma}} \tag{3-20}$$

This expression is convenient since it is independent of any specific value of antenna gain or size. It can be assumed for purposes of this study that the antenna gain and basic link are sized to produce a certain carrier-to-noise density C/N_0 in the receiver.

For an access control channel signal bit rate of about 1000 bps,[(3)] it may be assumed that the minimum C/N_0 to be received is about 40 dB-Hz during worst-case link conditions. This will produce a carrier-to-noise ratio, ρ_{Σ}, of 7 dB in an IF bandwidth of 2 kHz (a typical B_{IF} requirement for 1000 bps PSK). The sum channel (which is the communication channel)

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(3) 1970 DOT-CG Study, Op. Cit.

system noise temperature can be expected to be about 480 K (Table 2-1) for a transistor preamplifier with a noise figure of 3.5 dB. A worst case difference channel noise temperature might be twice this value (960 K) using less expensive preamplifiers with noise figures of about 6 dB. If it is further assumed that the monopulse error signal slope is such that the value of "m" is about 0.7, then from Equation (3-19),

$$\frac{\sigma_{\theta n}}{\sigma_{hp}} = 2.19 \times 10^{-2} \sqrt{B_\Delta} \qquad (3\text{-}21)$$

For typical values of position servo bandwidth, the tracking error is seen to be quite low. A typical servo bandwidth might be 5 Hz (two-sided) to accommodate the high angular accelerations experienced on ships. For this value, the rms tracking error due to noise would amount to 5% of the antenna beamwidth, causing in turn a very small (< 0.03 dB) effective gain loss.

b) Dynamic Lag Error

Actually, servo bandwidth is best selected for a minimum total error when combined with other bandwidth dependent errors.(35) For the MARSAT user application, the major bandwidth dependence is dynamic lag error. In typically second-order (double integration) autotracking systems the servo lag error is that resulting from acceleration inputs to the position servo loop. An acceleration input to the position servo causes a dynamic lag error of

$$\Delta\theta = \frac{\ddot{\theta}}{k_\alpha} \simeq \frac{\ddot{\theta}}{B_\Delta^2} \qquad (3\text{-}22)$$

where k_α is the tracking servo loop gain factor (acceleration error coefficient) which is numerically equal to the square of the servo loop

(35) Heckert, G.P. and Sordal, C.D. "Autotracking Accuracy of Large Antenna Systems for Satellite Communication Applications." Proc. of the IEE Conference on Steerable Aerials for Satellite Communications, Radio Astronomy and Radar. London, England. June 1966.

natural frequency in rad/sec. The natural frequency, ω_p, in rad/sec, can, in turn, be considered approximately equal to the two-sided noise bandwidth, B_Δ, of the servo loop, in Hz (the equality is exact if the servo system is represented by two poles and one zero with a damping factor of 0.5). The largest acceleration input is generally the azimuthal acceleration near the zenith. The use of a secant-correction device in the azimuth servo can reduce the azimuth error by the cosine of the elevation angle, i.e.,

$$\Delta A = \ddot{A}/(k_\alpha \sec E) = (\ddot{A} \cos E)/B_\Delta^2 \qquad (3\text{-}23)$$

The azimuthal error can be vectorially added with elevation angle error to yield total dynamic error with respect to the beam center, i.e.,

$$\Delta\theta_d = \sqrt{(\Delta E)^2 + (\Delta A)^2} = B_\Delta^{-2} (\ddot{E}^2 + \ddot{A}^2 \cos^2 E)^{\frac{1}{2}} \qquad (3\text{-}24)$$

This simple expression can be used, in conjunction with Equation (3-21) (or more generally Equation 3-19) to select a value of servo bandwidth which minimizes the total error. However, a value of antenna gain and beamwidth needs to be assumed. A high gain value of 14 dB corresponds to a beamwidth of about 33 degrees. Clearly, fairly large lag errors e.g., a few degrees, can readily be tolerated. If the same beamwidth is assumed in Equation (3-21), the three-sigma (peak) noise error becomes

$$3\,\sigma_{\theta n} = 2.16 \sqrt{B_\Delta} \qquad (3\text{-}25)$$

Assuming worst case angular rates as high as 45 deg/sec^2 (Section 2.4) the total dynamic lag error from Equation (3-24) can be approximated as

$$\Delta\theta_d = \frac{45}{B_\Delta^2} \qquad (3\text{-}26)$$

The peak total error, i.e., the sum of Equations (3-25) and (3-26), may be differentiated to show that it is minimized (to a value of about 7°) for a value of servo noise bandwidth of about 5.7 Hz.

Actually, the minimum total error as a function of servo bandwidth is a rather broad null, so that a lower value might be selected so as not to accept noise jitter most of the time just to keep dynamic lag error low during the small percentage of the time high dynamics occur. Since, for a 33 degree beamwidth, infrequent errors on the order of 1-2 degrees due to lag would be entirely acceptable, a value of position servo loop noise bandwidth of 4-5 Hz would be a credible selection.

c) <u>Wind-Gust Errors</u>

If the antenna is not mounted in an all-encompassing radome, wind can be a source of error. The wind force will be compensated by the tracking receiver and servo, but wind gusts can induce jitter. For the small antennas being considered here, wind gust effects on beam pointing should be negligible. Only in the maximum gain (15-18 dB) case might they be a consideration.

A spectral analysis of this effect can be made,[36] in which the (normalized) error spectrum due to wind gusts can be approximated as

$$S_e(\omega) = |\theta_e/T_W|^2 \; S_T(\omega) \tag{3-27}$$

where $S_T(\omega)$ is the wind torque spectrum and θ_e/T_W is the output angular error-to-input torque transfer function. The wind torque spectrum may be represented as shown in Figure 3-15, where σ_T is the standard deviation of the wind-gust torque, in lb.-ft., and ω_W is the wind-torque spectrum cutoff frequency, in rad/sec. The asymptotic approximation yields

$$\begin{aligned} S_T(\omega) &= \sigma_T^2/2\omega_W, & 0 < \omega < \omega_W \\ &= \sigma_T^2\omega_W/2\omega^2 & \omega_W < \omega < \infty \end{aligned} \tag{3-28}$$

The error/torque transfer function of a simple tracking antenna servo system

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(36) Newton, Gould and Kaiser, "Analytic Design of Linear Feedback Controls." John Wiley and Sons, Inc., New York, 1957.

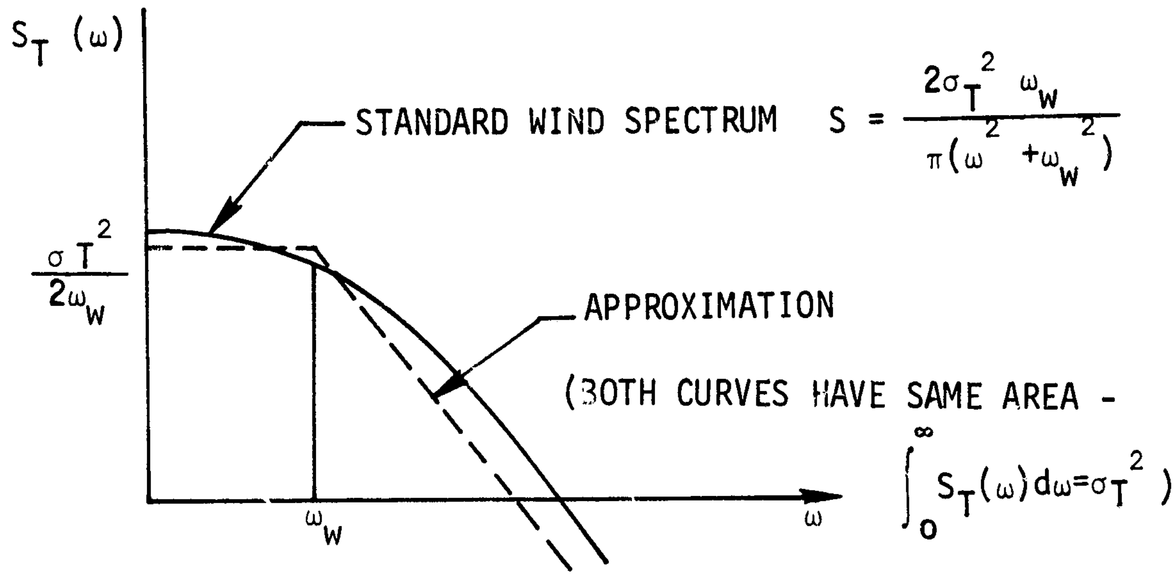

Figure 3-15. APPROXIMATION OF WIND-TORQUE SPECTRUM

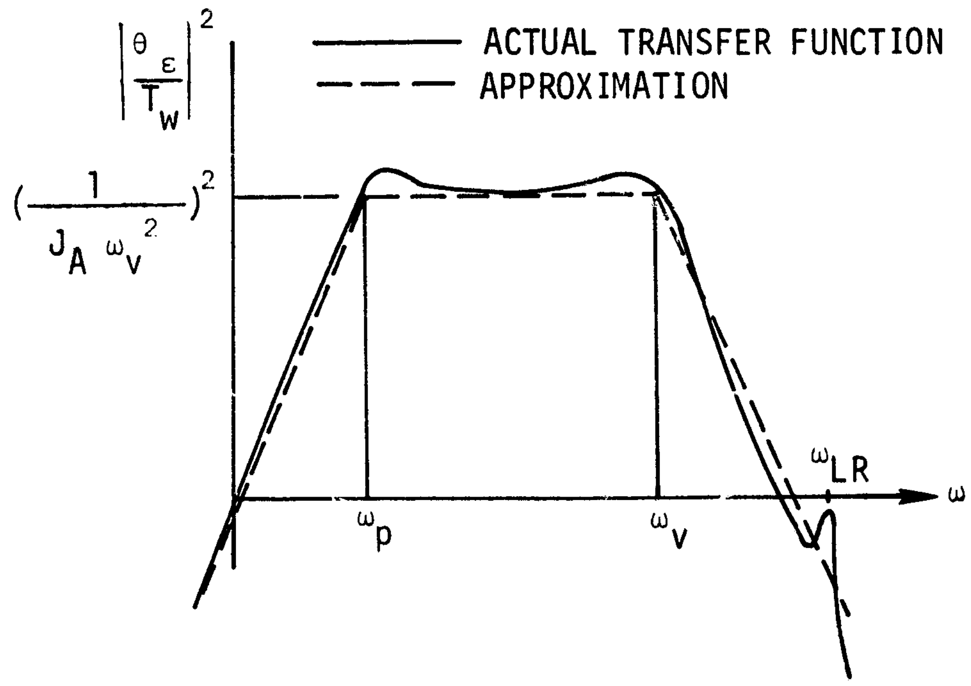

Figure 3-16. APPROXIMATION OF ERROR TORQUE TRANSFER FUNCTION

may be represented by the frequency response shown in Figure 3-16. The asymptotic approximation results in

$$\begin{aligned} |\theta_e/T_w|^2 &= (1/J_a\omega_v^2)^2\ (\omega/\omega_p)^4, && 0 < \omega < \omega_p \\ &= (1/J_a\omega_v^2)^2, && \omega_p < \omega < \omega_v \\ &= (1/J_a\omega_v^2)^2\ (\omega_v/\omega)^4, && \omega_v < \omega < \infty \end{aligned} \qquad (3\text{-}29)$$

where: J_a = antenna inertia, lb-ft-sec^2 (dyne-cm-sec^2)

ω_v = Velocity (tachometer) loop bandwidth, rad/sec.

ω_p = Position loop bandwidth, rad/sec

The mean-square error due to wind gusts may be found as the solution to

$$\sigma^2_{\theta_w} = \int_{-\infty}^{\infty} S_e(\omega)\ d\omega = \int_{-\infty}^{\infty} \left|\frac{e}{T_w}\right|^2 S_T(\omega)\ d\omega \qquad (3\text{-}30)$$

With the preceding asymptotic approximations and noting that $\omega_w < \omega_p = B_\Delta$, the above may be solved and reduced, yielding

$$\sigma^2_{\theta_w} \simeq [(57.3)^2/2]\ (\sigma_T/J_a\omega_v^2)^2\ (\omega_w/B_\Delta)\ \text{deg}^2 \qquad (3\text{-}31)$$

Examination of these parameters leads to the conclusion that wind-gust jitter can be made almost arbitrarily small for small antennas. The ratio of rms torque to inertia can be made small with appropriate load inertia in the motor. It is further reduced by the square of the velocity loop bandwidth (if there is a velocity loop) which must be larger than the position loop bandwidth. Finally, the wind torque spectrum cutoff ω_w is typically 0.1 rad/sec, which when divided by the position loop bandwidth of about 5 Hz, further reduces jitter.

d) Phase and Amplitude Unbalance

Gain unbalance between the feed lobes of a (simultaneous or sequential lobing) system induces a shift in the location of the tracking null. Gain unbalance will vary with incident signal polarization, but it is usually a fixed, bias-type error that can be controlled and calibrated out of the system.

In a monopulse tracking system, phase unbalance prior to the networks that derive the orthogonal error signals and the sum signal is termed pre-comparator phase shift, while relative phase variations between the reference (sum) signal and either of the error signals at the input to the error detectors are referred to as post-comparator phase shift. Simultaneous pre- and post-comparator phase shift will result in a shift of position of the tracking null with respect to the peak of the on-axis beam. The amount of pre-comparator phase shift at L-band can be kept small by proper design, so that post-comparator shifts have little effect.

Thus, the angle tracking error using conventional monopulse in the MARSAT application is comprised primarily of receive thermal noise and the angular dynamics associated with ships motion. As noted previously in part b) of this subsection, the total error due to these sources can be minimized by appropriate selection of the servo tracking loop bandwidth. It was shown that this value is about 5.7 Hz, assuming worst case ship dynamics, (45 deg/sec^2 combined acceleration input). However, since such accelerations occur only very rarely on small ships and essentially never on the large ships, a smaller value of servo position loop bandwidth would be more appropriate for design purposes. A value of 4 Hz is considered a conservative selection. It is optimum for acceleration inputs of about 15-20 deg/sec^2, (when the low received C/N_0 of 40 dB-Hz is included) and will limit the worst-case, small ship dynamic angular error to about 3 degrees, which is certainly tolerable.

Given the selection of 4 Hz for servo loop bandwidth, Equation (3-20) may be used to illustrate the performance of the conventional monopulse approach to the MARSAT user antenna pointing problem. This is presented in Figure 3-17. Two values of difference channel noise temperature are used in Figure 3-17; one in which the sum and difference channels employ the same low-noise preamplifier ($T_\Delta = T_\Sigma$) and one for the case of less advanced difference channel preamplifies, i.e., such that $T_\Delta = 2T_\Sigma$. (The sum channel system noise temperature is taken as about $T_\Sigma = 480$ K.)

Figure (3-17) also shows performance for two values of pre-detection IF bandwidth, 2 kHz and 4.8 kHz. As may be noted, this parameter has only a small effect on performance, as long as the signal-to-noise ratio (P_Σ) is greater than unity (Equation 3-15). These values of bandwidth are used to illustrate two possible access-control channel signals, upon which the angle-tracking may be based. The 2 kHz case assumes the channel contains only the access-control signal, for which about 40 dB-Hz is needed at the receiver, and the second case, $B_{IF} = 4.8$ kHz, assumes that the access-control message of 1000 bps is multiplexed with Telex or other working data channels, comprising about 1400 bps, for a total channel bit rate of 2400 bps, for which about 45 dB-Hz is required at the receiver.

It may be readily seen that the angle-tracking error (normalized by the half-power antenna beamwidth) and its corresponding effect on operational antenna gain is very small, i.e., 0.01 dB or less. Clearly, this is insignificant. It may be concluded, therefore, that monopulse angle tracking would perform very well in the subject application.

However, multichannel monopulse receivers do have to be aligned in phase from time to time, hence this is a disadvantage of the conventional monopulse approach. The most significant disadvantage is, however, that it requires two additional receiver channels in the terminal.

In this context, it is possible to reduce the number of separate channels required from three to two by combining the two difference signals in phase

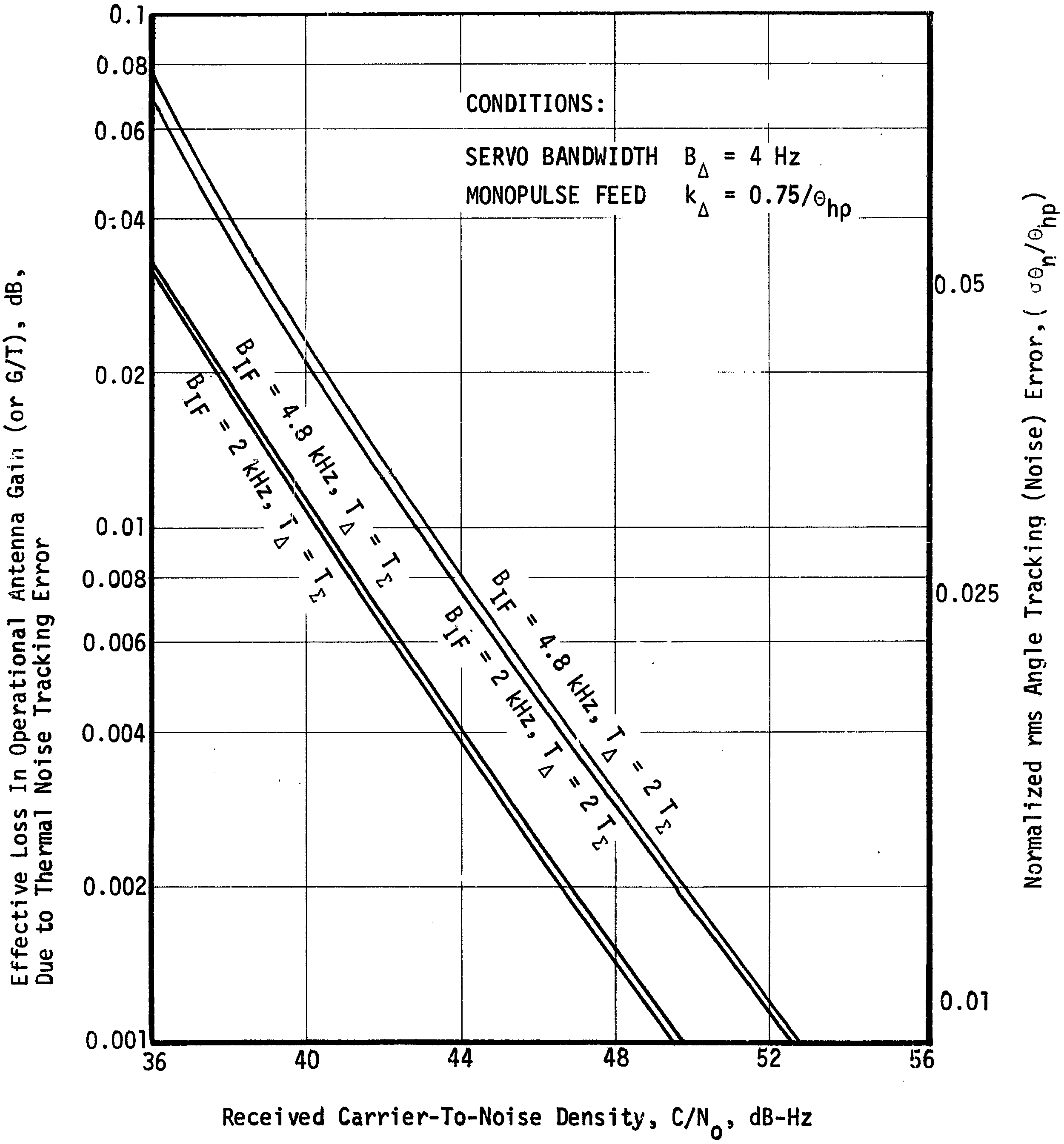

Figure 3-17 MONOPULSE ANGLE TRACKING PERFORMANCE

quadrature into a single complex signal. The complex difference signal can then be amplified and filtered in a single channel, after which the two original difference signals can be separated at IF by resolving the complex signal into its components in-phase and 90° out-of-phase with respect to the sum signal. This technique reduces the number of difference channels required but does so at the expense of tighter phase tolerances. The phase alignment problem and additional circuitry disadvantages can be alleviated almost entirely by the use of a single channel monopulse approach. This concept is considered next.

3.5.3 Single Channel Monopulse (Monoscan)

The single channel monopulse approach uses a monopulse antenna with a conventional comparator circuit, but then combines the azimuth and elevation error signals into a single complex error signal, which in turn is used to amplitude modulate the sum channel signal. This combining is all done at RF prior to amplification, so that only a single receive channel is required. This advantage is gained at the expense of the insertion loss of the combiner in the communication signal path, and a reduced tracking accuracy relative to conventional multichannel monopulse. It is a very practical approach when simplicity is required and a small degradation in user terminal G/T is permitted.

3.5.3.1 Configuration Factors

Figure 3-18 is a block diagram of the user terminal receiver and access control channel demodulator equivalent to that of Figure 3-14, except with the required additions shown for a single instead of three channel monopulse approach. Because the azimuth and elevation error signals are (after being combined) amplitude modulated on the sum channel signal, the receiver function is equivalent to that of a conical-scan detector, and thus single channel monopulse is referred to as psuedo conscan, and

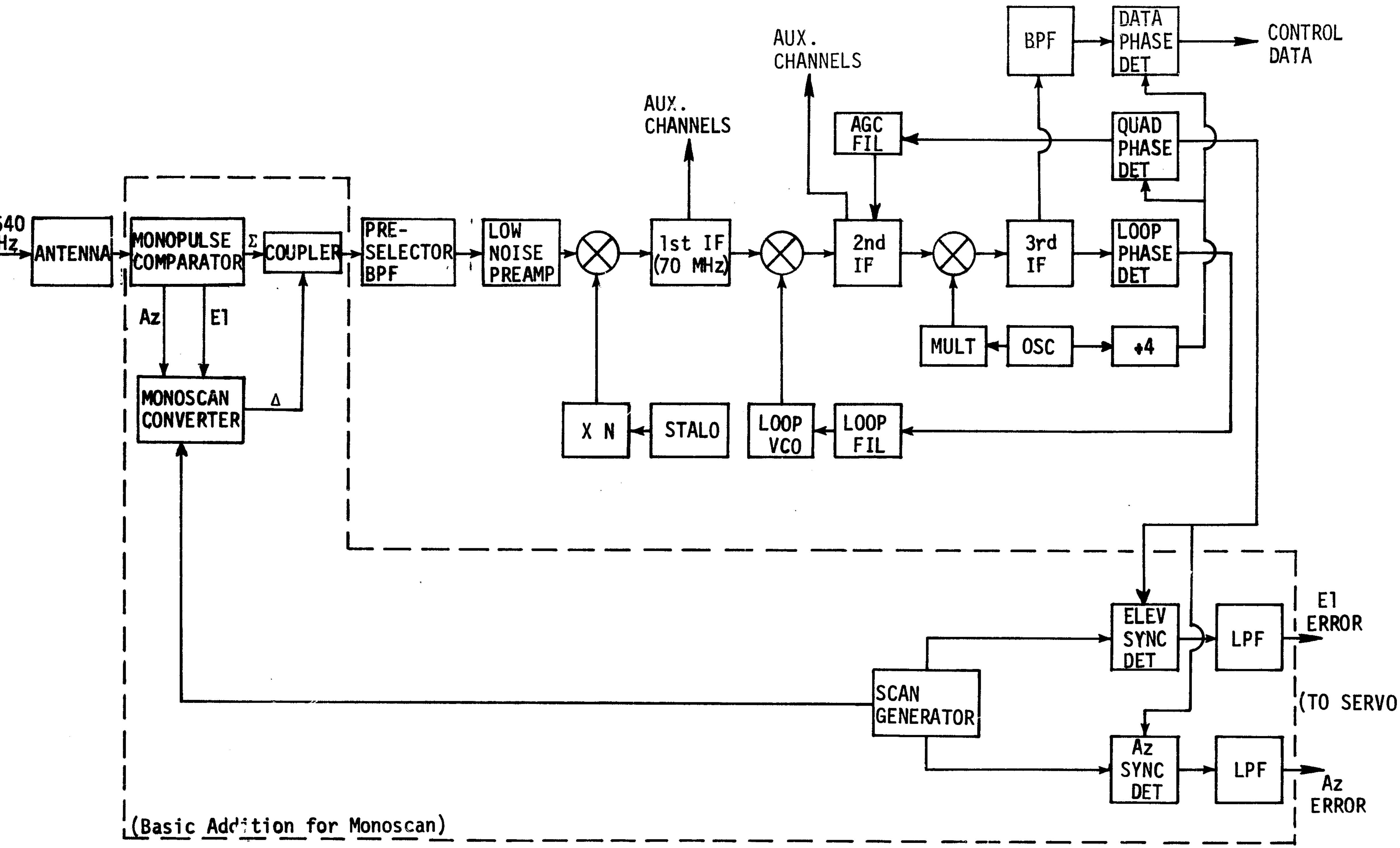

Figure 3-18 SINGLE-CHANNEL MONOPULSE (MONOSCAN) EXAMPLE

more recently monoscan. The coupler, by simply permitting the addition (and subtraction) of signals, acts as an AM modulator.

The AM modulation index is a function of the monopulse (feed) pattern error slope (k_Δ) as well as the coupling factor. Increasing coupling of the combined difference signal increases modulation index and thus angle-tracking accuracy, but it does so at the expense of increased coupler insertion loss. Since the coupler precedes the low noise preamplifier in the communication receiver, the insertion loss affects antenna gain directly, and also increases system noise temperature. The usual design goal is to select the modulation index as low as possible consistent with tracking accuracy and AM detectability requirements.

As may be noted in Figure 3-18, the receiver's AGC detector serves the AM detection function. Its video output is routed to an azimuth and elevation detector, which is synchronous with the scanning function used to combine the two channels at RF. Historically, the difference channels were combined in phase-quadrature (e.g., by motorized scanners), and then separated by use of quadrature angle error detectors. Current technology permits the combining, scanning and synchronous detection to be implemented digitally. Because monoscan is susceptible to interference at the scan rate (as in mechanical conscan), the scan rate may be randomly coded to prevent degradation from interference, as shown by Pifer.[37] The combining function can be implemented by the switching of the difference signals through fixed phase-shifters, such as latching phase shifters, in an appropriate sequence.

A functional block diagram of the digital monoscan combiner is shown in Figure 3-19. The switching sequence is periodic and driven by the "scan

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(37) Pifer, P.M. "Technical Discussion - Scan Coded Single Channel Monopulse Automatic Tracking Systems." Scientific-Atlantic, Inc. Report. Atlanta, Georgia. 10 January 1967.

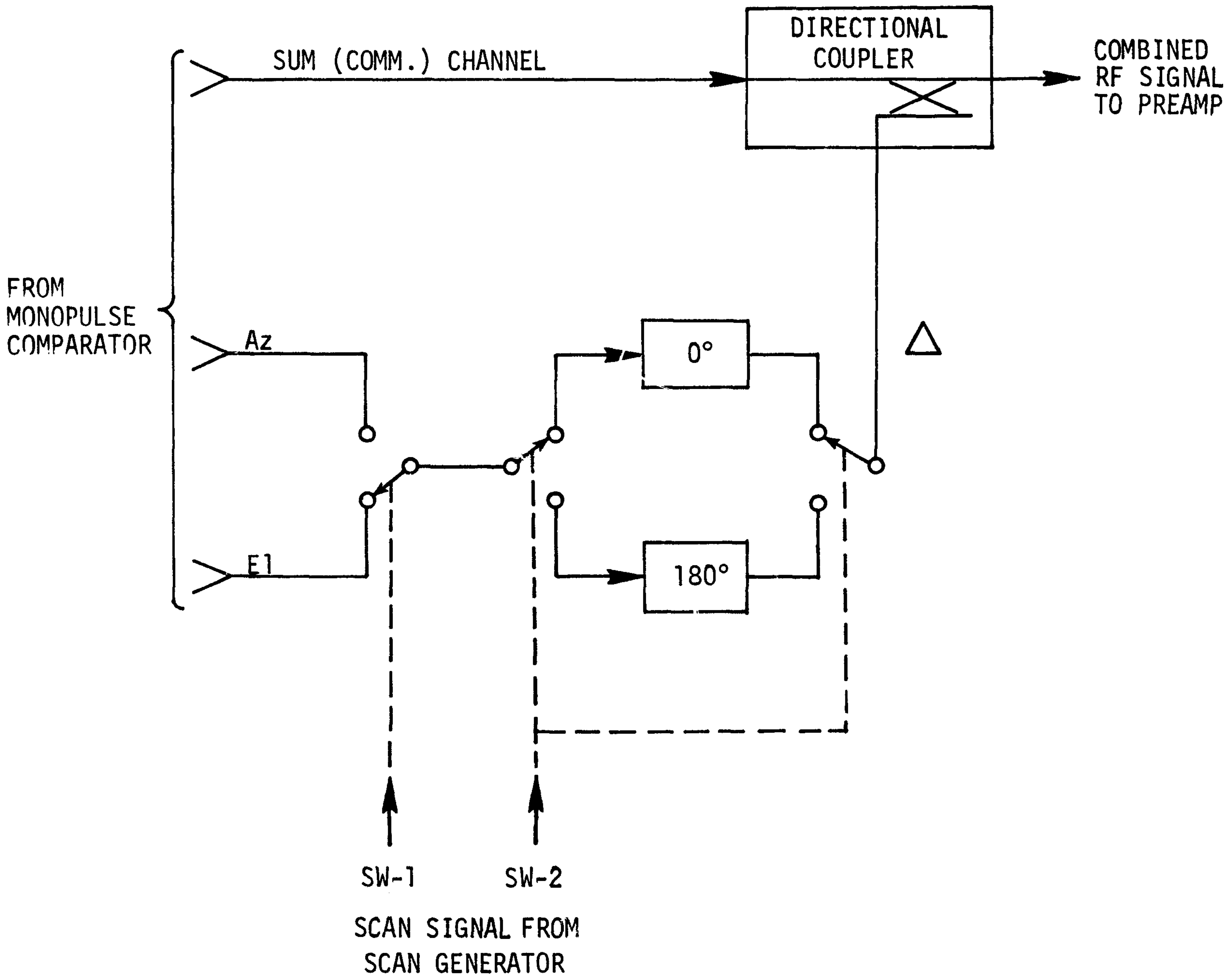

Figure 3-19 MONOSCAN CONVERTER FUNCTIONAL BLOCK DIAGRAM

generator" shown in Figure 3-17. The scan rate can be very low, e.g., 100 200 Hz. It must be selected to be consistent with AGC time constant (so that AGC does not "see" the AM) and tracking bandwidth. The synchronous video detectors used to produce dc azimuth and elevation error signals can be gated phase detectors, driven by the same timing signal controlling the RF switches SW-1 and SW-2.

The monoscan approach is being used more and more in satcom terminal applications, especially military. Its primary advantage is reliability and simplicity of operations, in that no phase alignment procedures are required. The advantages of significantly reduced circuitry are also important relative to conventional monopulse in the MARSAT application.

3.5.3.2 Relative Tracking Accuracy of Monoscan

As with conventional monopulse, the accuracy of a monoscan tracking system is limited by thermal noise errors, dynamic lag and wind-gust jitter. Errors due to intra-channel phase shift or gain unbalance, however, do not exist, especially in the narrow-band application considered here. The dynamic lag pointing errors and wind-gust jitter is of course identical for the two tracking receiver configurations.

The thermal noise tracking error, however, is different. An analogous expression to Equation (3-19) and (3-20) for monopulse can be derived for monoscan (Appendix C). The result is

$$\sigma_{\theta_n}^{2} = \frac{1}{k_m^2} \left(\frac{1 + 2\rho_\Sigma}{\rho_\Sigma^2} \right) \frac{B_\Delta}{B_{IF}} \qquad (3\text{-}32)$$

where k_m is the monoscan (amplitude) modulation index, and the other parameters are as for monopulse. The modulation index is given by

$$k_m^2 = k_\Delta^2 \frac{1}{L_c (L-1)} \tag{3-33}$$

where L is the coupling factor used to insert the combined difference channel signal on to the sum channel, and L_c is the Az-El combiner (multiplexer) line loss. Upon appropriate combination of Equations (3-33) and (3-18), with (3-6) and (3-32), the effective gain loss for monoscan may be shown to be

$$L_t \text{ (dB)} = \frac{12}{m^2} L_c (L-1) \left(\frac{1 + 2\rho_\Sigma}{\rho_\Sigma^2} \right) \frac{B_\Delta}{B_{IF}} \tag{3-34}$$

Upon division of Equation (3-34) by Equation (3-20) to obtain a ratio of pointing offset losses (in units of dB) as a comparison of monoscan (ms) and monopulse (mp), the following results:

$$\frac{L_t \text{ (dB) ms}}{L_t \text{ (dB) mp}} = \frac{L_c (L-1) (1 + 2\rho_\Sigma)}{(1 + \rho_\Sigma) T_\Delta/T_\Sigma} \tag{3-35}$$

At high input signal-to-noise ratios, $\rho_\Sigma >> 1$, and assuming low combiner loss L_c, the above ratio can be approximated as

$$\frac{L_t \text{ (dB) ms}}{L_t \text{ (dB) mp}} \sim \frac{2L}{T_\Delta/T_\Sigma} \tag{3-36}$$

In a typical monoscan system, the coupling factor L should be at least 10 dB, and for a monopulse system which, as noted in Section 3.5.2.2 a) to be applicable does not employ special low-noise difference channel preamplification so that $T_\Delta/T_\Sigma \simeq 2$, the ratio of losses is L. This is typically a factor of about 10, which is clearly quite high. It would imply that monoscan is only suitable when the equivalent monopulse noise error is very small. In the preceding monopulse example, it was shown that the rms angle tracking error due to noise would be about 5% of the half-power beamwidth. From Equation (3-20) (or Figure 3-17) the effective gain loss of the monopulse antenna would be about 0.02 dB. This means that the loss in gain when using monoscan with the same antenna would be about 10 times this, i.e., 0.2 dB.

The general tracking performance of monoscan is shown in Figure 3-20, assuming a 4 Hz servo bandwidth for purposes of comparison with Figure 3-17, for various monoscan coupling factors. The effect of IF bandwidth is also indicated by the dashed-line. The difference between the two cases, 2.0 kHz and 4.8 kHz, would be significant only at very low input signal levels. In the region of interest, 40-45 dB-Hz, the error would be between 7% and 25% of the half-power beamwidth. Clearly, low values of coupling L are desirable. However, while low values of monoscan coupling factor L reduce the tracking error due to receiver thermal noise, they produce an increased constant degradation due to line loss in the primary communications channel.

In general, the pointing error with monoscan can be reduced by increasing AM modulation index. This is done by lowering the coupling factor, previously assumed to be 10 dB. However, this increases the insertion loss in the communication channel, and thus G/T directly. The insertion loss of the monoscan coupler in the sum channel is

$$L_{\Sigma} = \frac{L}{L - 1} \tag{3-37}$$

For a 10 dB coupler, the insertion loss is about 0.45 dB. When slight resistive losses are included, its value would be about 0.5 dB. This loss subtracts directly from antenna (receive) gain, and further adds about 30 K to the system noise temperature, for a net degradation in G/T of almost 0.7 dB.

The addition to receive system noise temperature is given by

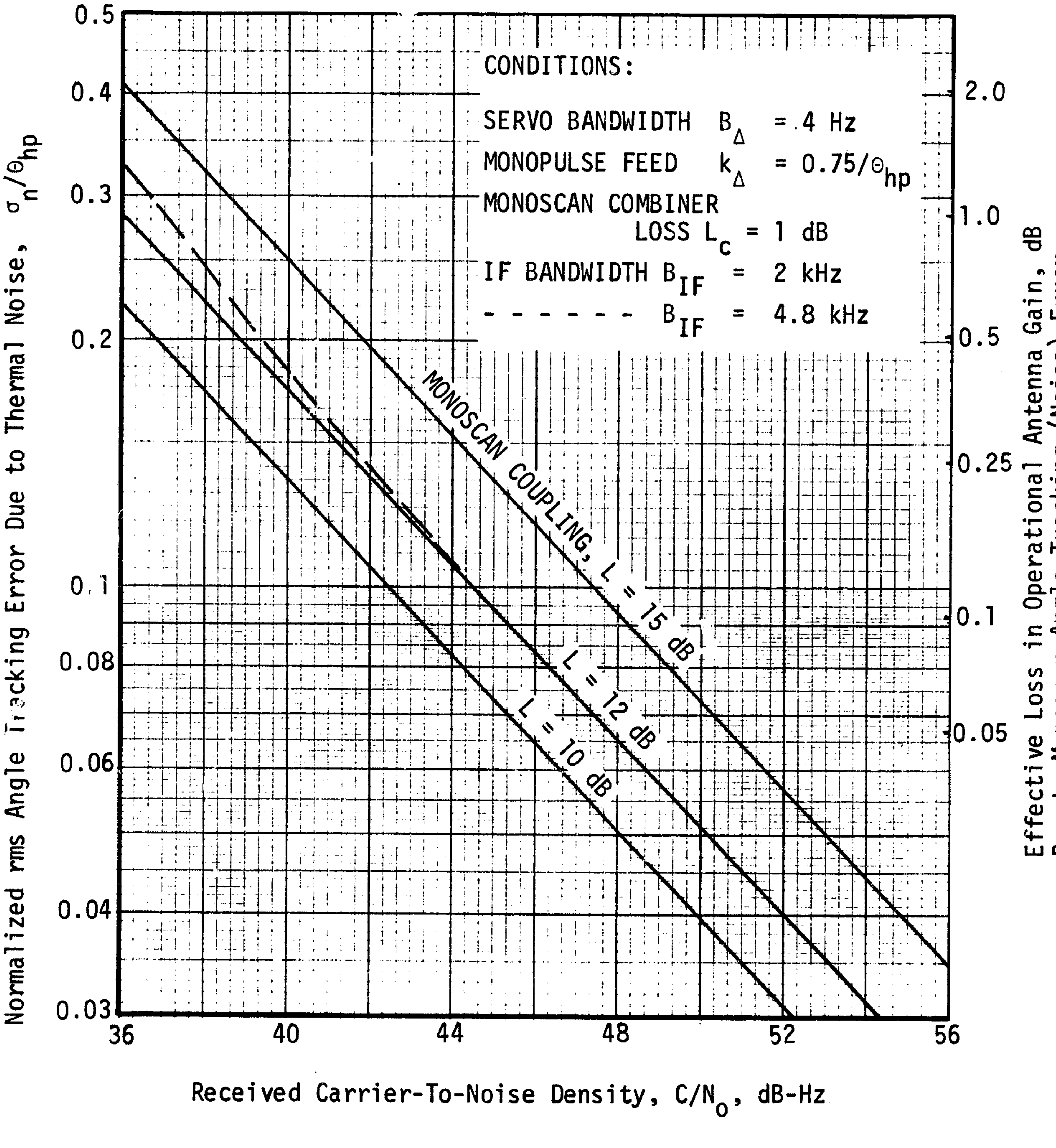

Figure 3-20 MONOSCAN ANGLE TRACKING ERROR DUE TO RECEIVER THERMAL NOISE

$$\Delta T_s = \frac{(L_\Sigma - 1)\, T_0}{L_\Sigma} = \frac{T_0}{L} \tag{3-38}$$

where T_0 is the ambient temperature, e.g., 290 K. (Figure C-3 in Appendix C depicts the general effect of monoscan coupling factor on antenna gain and system noise temperature.)

When the effects of coupling of the combined difference channel on to the operational, average antenna gain degradation due to pointing error, a total effective loss in system receive performance can be ascribed to the monoscan is illustrated in Figure 3-21. This net degradation is then comparable to that of an "ideal" tracking subsystem, e.g., the conventional monopulse performance shown in Figure 3-17. Clearly, selection of the coupling factor must by carefully considered. While it is indicated in Figure 3-20 that a low value of coupling L is desirable, Figure 3-21 indicates that overall, a higher value is better, at least for high received carrier-to-noise density ratios. At 40 dB-Hz, the value of L = 12 dB is superior. Above 42 dB-Hz, it appears that L = 15 dB is superior.

When the coupling is 15 dB, however, a large portion of the loss is due to receiver thermal noise. For example, at 45 dB-Hz, received C/N_0, the rms tracking jitter (Figure 3-20) is 13.5% of the half-power beamwidth of the antenna. For a peak gain of 12 dB, the beamwidth, Θ_{hp} = 40 deg, which would then imply an (rms) angle jitter of ±5.4 deg. Instantaneous peaks, corresponding to the three-sigma probability, could cause fluctuations as high as ±16 deg. It would be desirable to reduce this jitter, even at the expense of an increased constant degradation in effective antenna gain.

For the subject application, where aperture efficiency is by no means critical, it would appear judicious to select a compromise value of coupling factor such as L = 12 dB. Relative to 15 dB, this effectively

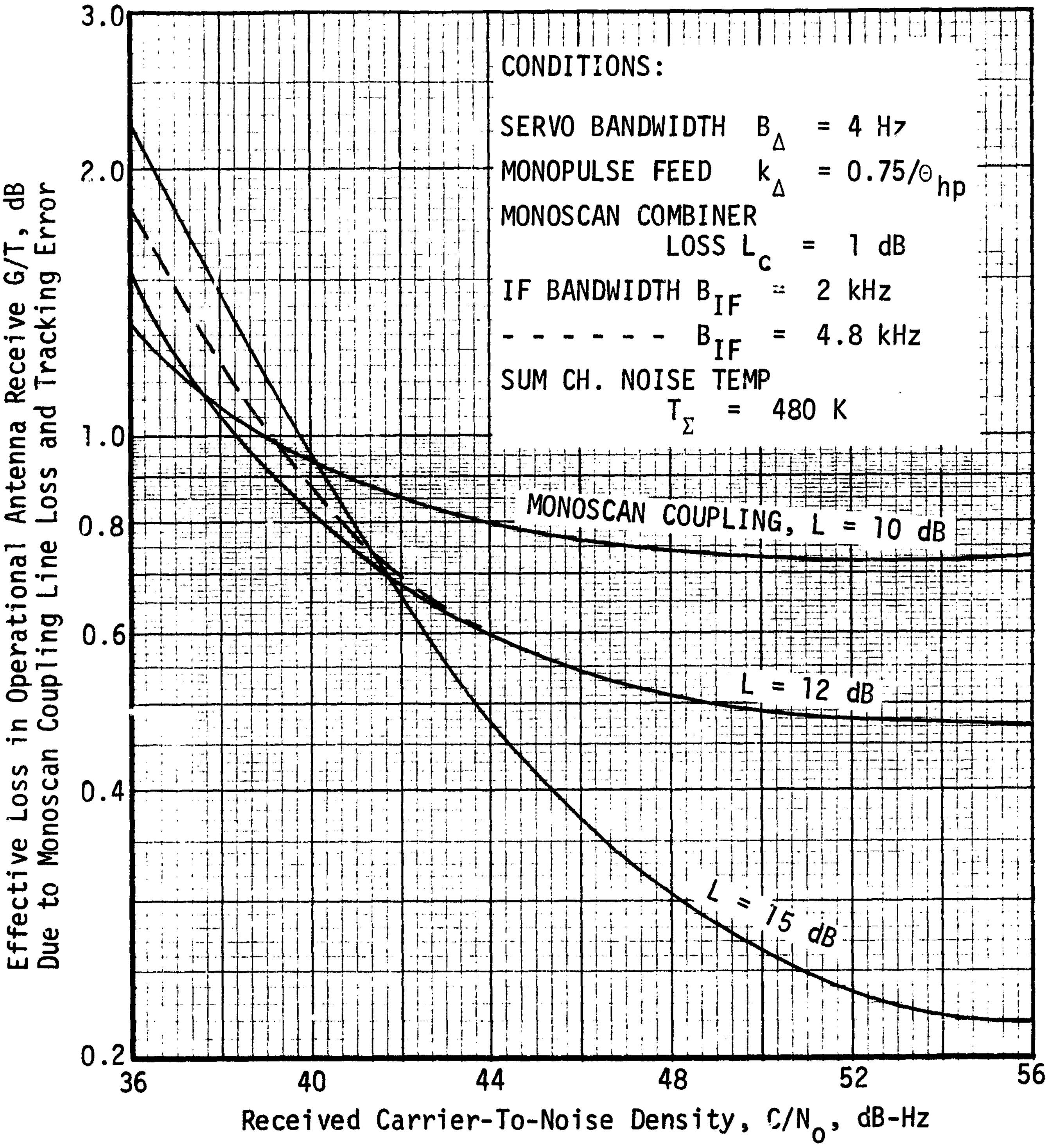

Figure 3-21 NET PERFORMANCE OF MONOSCAN TRACKING

degrades antenna efficiency by 0.26 dB, which corresponds to 3% of antenna aperture (equivalent) diameter. For a short-backfire antenna with a diameter of 39 cm (15.4 inches), this corresponds to 1.2 cm (0.46 inches) of "wasted" diameter. Surely, this is a small penalty to pay to decrease the rms jitter from 13.5% of the beamwidth to 9.4% (about ±3.7 deg). The conclusion to this trade-off is then that for monoscan tracking in the subject application, a sum channel coupling factor of 12 dB, and perhaps even 10 dB (depending on radiator and peak gain selection) is most suitable.

It may be concluded, therefore, that use of the monoscan technique would typically produce a degradation in effective antenna gain, or efficiency, of about 0.8 dB to 0.54 dB, depending on whether 40 dB-Hz or 45 dB-Hz is used as the design value of received C/N_0 for angle tracking, i.e., in the access-control (data) channel receiver. This degradation is quite small for the operational reliability associated with the single-channel monopulse automatic tracking technique.

3.5.4 Comparison of Autotracking Schemes

Section 3.5 has contained a summary of the analysis of performance of conventional automatic angle tracking schemes for shipboard antennas operating at L-band. The schemes are "conventional" in that the antenna incorporates a feed arrangement amenable to the formation of a sharp pattern null on the boresight axis. This permits an appropriately designed receiver to seek and lock-to the antenna pattern null. The signal which is tracked can be any one or all of the signals transmitted by the satellite. Clearly, the best signal is one which is transmitted by the satellite at all times and at a fixed frequency. In as much as a special non-working channel is needed for system (demand) access control, this signal is considered most suitable as a tracking beacon.

It was noted that conical-scan angle tracking schemes employing mechanical rotation of a feed or antenna are generally undesirable for the subject application due to reliability, accuracy and the fact that the shipboard transmitted signals would be amplitude modulated, albeit by a small amount. The trade-off between performance and complexity is then between various versions of monopulse angle tracking. These are distinguished primarily by the number of channels used to implement the tracking receiver function.

Figure 3-22 contains an overall performance comparison of single channel, two-channel and three-channel monopulse tracking. The performance is expressed in terms of the effective degradation in the antenna subsystem's receive gain-to-system noise temperature ratio G/T that would occur in operation, relative to an ideal angle tracking system. For two- and three-channel monopulse systems, the effective degradation is due to the fact that there is an average pointing error due primarily to receiver thermal noise. On the average then, the desired satellite signal is not being received at the peak of the antenna beam, but at a slight offset. As seen in Figure 3-22, for three-channel monopulse, this degradation is negligibly small over the range of received C/N_0's of concern. Two-channel monopulse, in theory, is only slightly worse, i.e., by about 25%, due to the effect of difference channel combiner loss (about 1 dB) on the slope of the error pattern.

For single-channel monopulse, the effective degradation in user terminal G/T is considerably higher. In this case, it is due to a fixed line loss associated with insertion of a coupler in the sum channel, prior to preamplification, as well as angle pointing error. The fixed line loss is about 0.3 dB, which decreases effective gain by that amount, as well as increasing the system noise temperature by about 18 K, which corresponds approximately to about 0.16 dB (assuming a nominal 480 K system noise temperature). Thus, the total fixed loss in G/ T is about

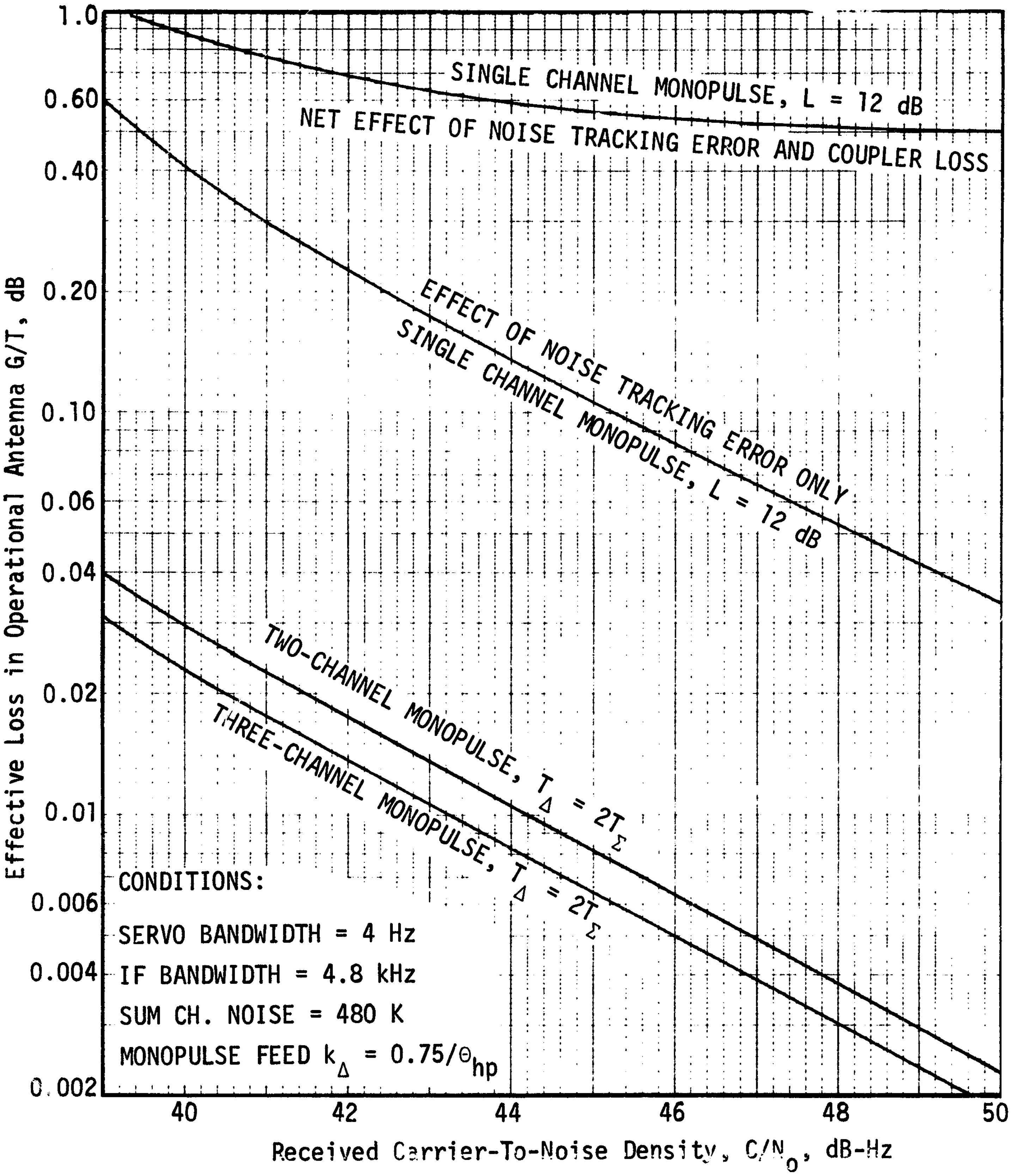

Figure 3-22 COMPARISON OF SINGLE-CHANNEL AND MULTI-CHANNEL MONOPULSE PERFORMANCE

0.46 dB, for the 12 dB sum channel coupling factor considered most suitable (Section 3.5.3).

The angular tracking error itself for single-channel monopulse is also considerably larger than that for the multichannel versions. It is larger than for three-channel monopulse by a factor of $\sqrt{L}$, i.e., about four. Its effect on G/T by itself, however, is only 0.11 dB, assuming a received C/N_0 of 45 dB-Hz in the access control/data channel. This corresponds to an rms noise tracking error of about 9.4% of the antenna's half-power beamwidth. This amount of angular error would exist only the small percentage of the time that the above C/N_0 exists. That is, if 45 dB-Hz is the near-worst case design value for received C/N_0, most of the time the signal levels will be higher than this, with correspondingly smaller errors.

The above angle tracking error can also be reduced by decreasing the servo position loop bandwidth below the 4 Hz value assumed in Figure 3-22. This value is strictly optimum only when the angular acceleration input (roll, etc.) is 45 deg/sec^2, a very conservative design point even for small ships. For example, using the worst-case angular acceleration of 7.5 deg/sec^2 for large ships (Section 2.4.2), the theoretical optimum value of servo bandwidth is 2 Hz. This value would reduce the angular noise error from 9.4% to 6.6% of the antenna beamwidth, and thus decrease the effect on G/T from 0.11 dB to less than 0.06 dB.

As these angular errors are modest, it is concluded that while the accuracy of multichannel monopulse is clearly superior, the accuracy of single-channel monopulse is adequate for the subject shipboard application. Table 3-2 presents an overall companion of the basic techniques considered.

The single-channel technique is the simplest to implement and, consequently, the most reliable and economical. It only requires one receiver

TABLE 3-2

TRACKING TECHNIQUE COMPARISON

Trade-off Parameter	BASIC TECHNIQUE		
	Single-Channel Monopulse (Monoscan)	Two-Channel Monopulse	Three-Channel Monopulse
Tracking Accuracy	Adequate; tracking error due to thermal noise approx. 4 times larger than 3-channel monopulse	Good; tracking error due to thermal noise approx. 12% larger than 3-channel monopulse	Very Good
Direct Effect on G/T	Reduced by approx. 0.5 dB due to coupler	No reduction in sum channel gain	No reduction in sum channel gain
Complexity	Simplest	Complex	Complex
Number of Receiver Channels	One	Two	Three
Reliability	Most reliable	Fair	Fair
Phase Adjustment over Receive Band	Not required	Required	Required
Potential Problems	AM interference; (can be solved by pseudo-random sampling technique)	None	None
Development Risk	None	None	None

channel. Since this channel is needed also for the primary communications functions, it can be said that the single-channel, or monoscan approach requires no additional receiver channels. Its main functional advantage is that it requires no difference channel phase-shifter adjustment when the channel frequency being tracked is changed. This is due to the fact that there are no active elements before the sum and difference channels are combined.

The two-channel and the three-channel systems are more complicated to implement. They require additional total receiver channels. Because of the additional channels, they are also less reliable, harder to maintain, and more costly. For proper demodulation (cross-correlation), the differential phase shift between the sum and the difference channels at the cross-correlator must be kept small ($\sim 20°$). Since each channel is amplified by a separate chain, changing carrier frequency would require readjusting the phase shifters. (Control channel carrier frequency need be changed only when switching between satellites.)

It is concluded therefore, that of the various conventional automatic tracking techniques, monoscan is preferrable for the MARSAT user application. The cost differential between the techniques is highly dependent on the type of antenna used, but is at least $1000 when just the additional preamplifiers, filters, down-convertors, transmission lines, IF channels, etc. are considered. It should finally be re-noted that the effect of the slight degradation of 0.5 dB, for example, in receive G/T due to insertion of a monoscan coupler can be compensated readily by an additional 6% in (equivalent) aperature diameter, which, for the gains being considered at 1600 MHz, is on the order of only 2.5 cm, which is hardly significant.

3.6 MECHANIZED MANUAL TRACKING - THE STEP-TRACK SCHEME

While conventional satellite signal automatic angle tracking schemes have certain undesirable features, i.e., a tracking feed or equivalent RF circuiting for forming of difference patterns, the manual tracking function is always possible as long as the antenna beam is pointable. As described in Section 3.3, this function is performed by a (radio) operator directing (or selecting) the antenna beam so as to maximize received signal level.

It is not fundamental that a human operator is required for this function. Simple analog or digital circuitry could be used to provide the necessary interface between existing signal demodulators and the antenna beam position controller. Such a scheme has been proposed, primarily for applications such as tracking of synchronous communication satellites by fixed earth terminals wherein angular dynamics are extremely small.[38]

A computer could be used to perform the operator's functions, including the provisions for rate and direction memory. However, use of a computer would not be consistent with the goal for MARSAT user terminal simplicity. The more simple form of "automated" or mechanized manual tracking suggested is referred to as "Step-Track," as the antenna is made to Step-Towards-the (received)-Energy-Peak. Only analog circuitry is used and, as such, the scheme does not readily provide for rate or direction memory. In this mechanization, the antenna "scanning" would be continuous, e.g., in small fixed angular steps (in both axes), such as readily provided for by stepper-motor type drives. Without the computer, no rate or direction memory would be practical. The D.C. voltage input received could come from any demodulator (such as the access control modem) with AGC or S-meter monitors, or from a separate IF energy detector preceded by two or three

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(38) Tom, N. N. and Heckert, G.P., "Step-Track - A Simple Autotracking Scheme for Satellite Communication Terminals," Paper 70-416 Presented at the AIAA 3rd Communications Satellite Systems Conference, Los Angeles, Calif. April 6-8, 1970.

selectable bandpass filters, carefully chosen for their applicability to "specific" communication spectra. Such detection possibilities are indicated in the functional block diagram of the Step-Track scheme in Figure 2-23. While the concept is applicable to electronically-steered phased arrays and generally to switched beam antennas, it is convenient to consider its functional characteristics and performance in the context of a mechanically driven antenna pedestal.

The Step-Track scheme is relatively simple to implement and attractive for certain classes of terminals. Its suitability is questionable for the high angular dynamics possible with ships, and for this reason must be analyzed for the MARSAT user application. Further, it must be noted at the onset that any form of manual tracking, automated or not, has a number of limitations. It is almost axiomatic that locating a beam maximum can never be as accurate as finding a sharp null. Further, manual tracking can be degraded by fluctuation in the received signal levels owing to atmospheric perturbation, or satellite antenna stabilization. As a result, the effective gain of the antenna will be reduced slightly. For example, a pointing error of 0.18 beamwidth will cause a loss of 0.4 dB in gain. Such an error would be attendant with only 5% uncertainty in AGC voltage. However, this degradation in gain may be partially offset by the use of a more efficient nontracking feed relative to the typical tracking feed and comparator assembly.

Thus, while the Step-Track scheme can be expected to be less accurate than conventional autotracking techniques, it constitutes what might be considered the ultimate in simplicity. It is most cost-effective for certain classes of ground terminals. Its use could delete the requirements for a tracking feed, microwave comparator circuitry, receiver difference channels, and the conventional angle tracking receiver. As substitutes, certain antenna drive servo logic and interface with the communication/data receiver would be required.

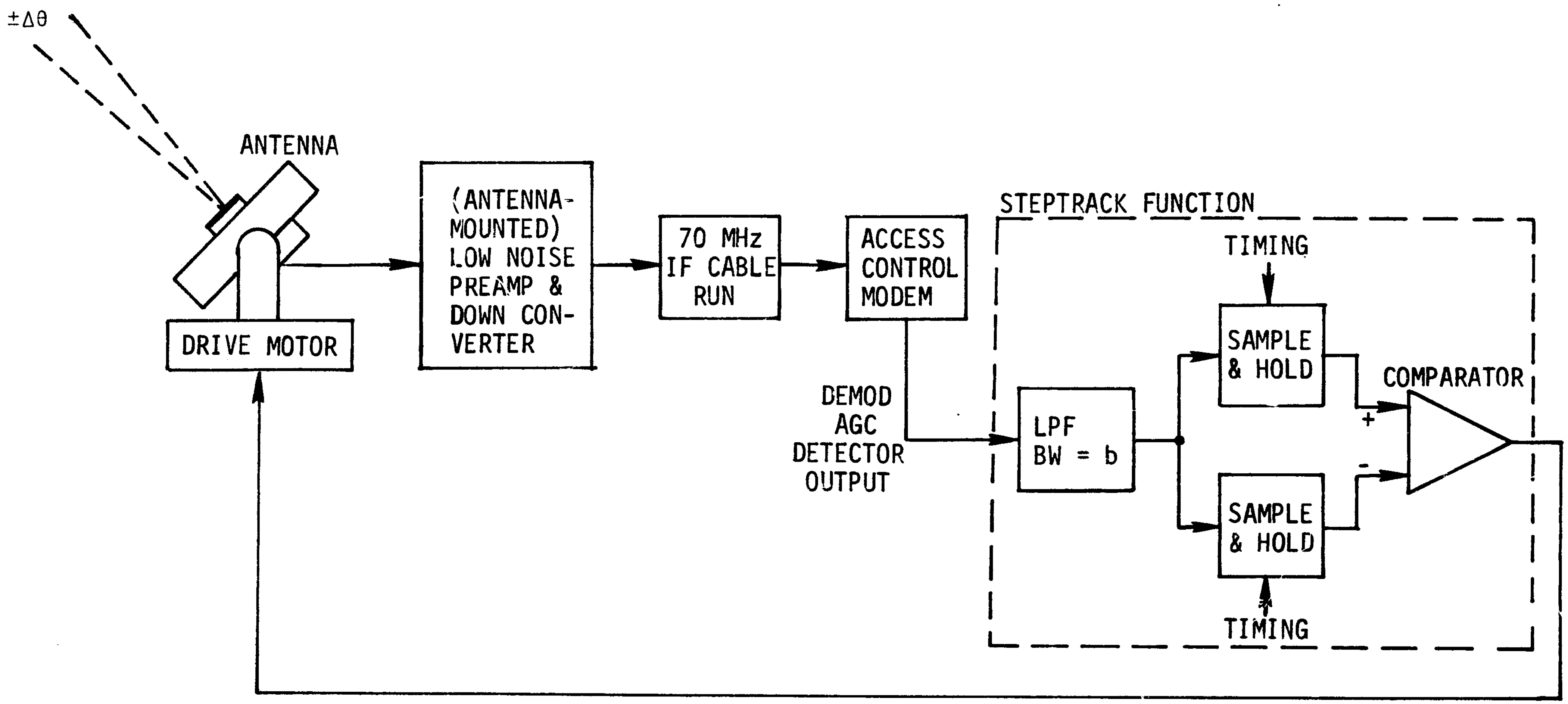

Figure 3-23 FUNCTIONAL MODEL OF STEP-TRACK CONCEPT

3.6.1 Operation and Implementation of Step-Track

A general functional diagram of the Step-Track concept depicting major functions and, in particular, potential interface with other subsystems of the terminal, is shown in Figure 3-23. The dc voltage input into the decision circuit, representing the received signal level, could come from any demodulator equipped with AGC or S-meter monitors. The decision circuit required and its associated timing generator are shown in Figures 3-24 and 3-25, respectively.

The decision circuit consists of two sample/hold and integrator modules, a voltage comparator, and logic gates. The two sample/hold and integrator modules are used to sample the input voltage levels before and after the antenna has been moved by a preset increment, $\Delta\theta$. Then, these two samples are compared. If the amplitude is increasing, as given by the value of the comparator output, the antenna is moved in the same direction the next time that same axis is to be moved. If the amplitude is decreasing, the direction of movement would be reversed. This process would be continuous, with the movement alternating between the X and Y axes. The control signals to the servo drives of the two axes are regulated by the logic gates.

The timing generator, consisting of an oscillator, a counter, and output gates, generates four sequences of control pulses for the decision circuits. The time relationships of these pulse sequences are shown in Figure 3-26. The pulses in the first sequence control the sampling operation of the Sample/Hold and Integrator Module #1. The second sequence operates the gate for the command signal to the X-axis stepper motor. The third sequence operates the Sample/Hold and Integrator Module #2, and the fourth sequence controls the command signals to the Y-axis stepper motor.

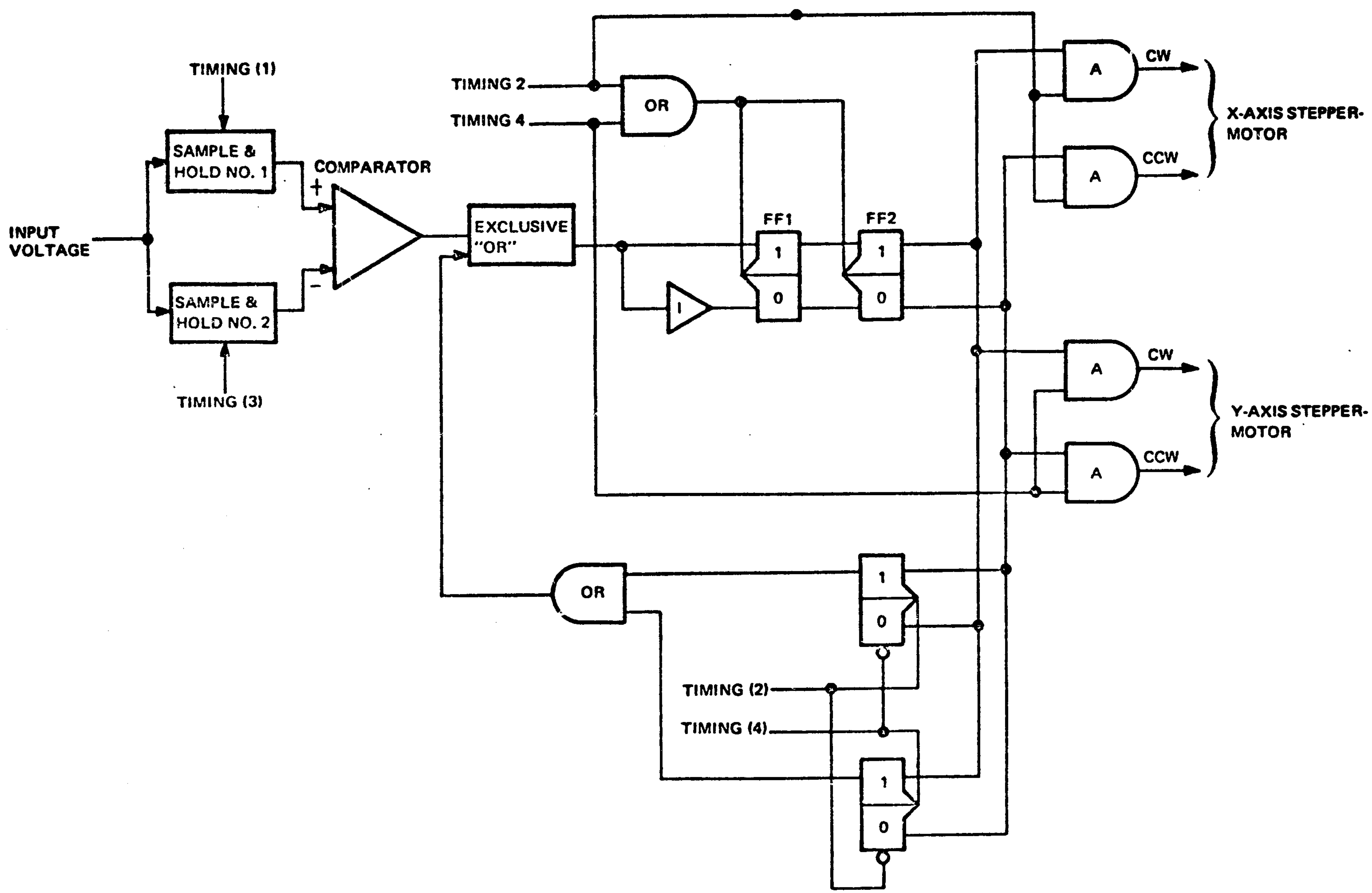

Figure 3-24 STEP-TRACK DECISION CIRCUIT

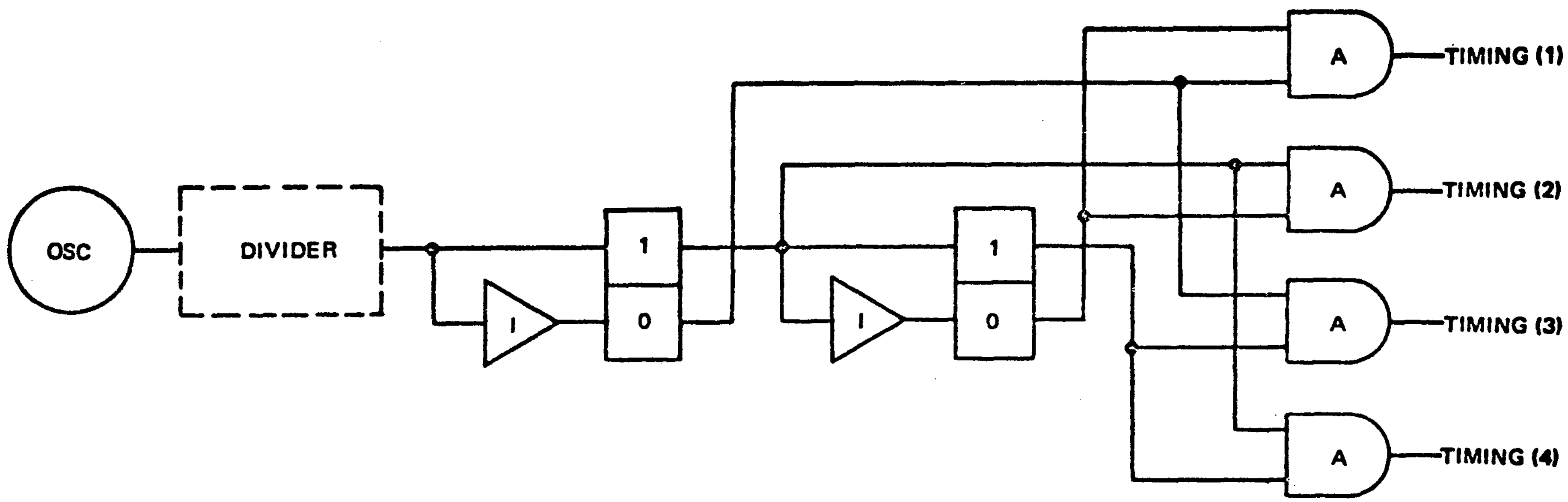

Figure 3-25 TIMING GENERATOR FOR STEP-TRACK LOGIC

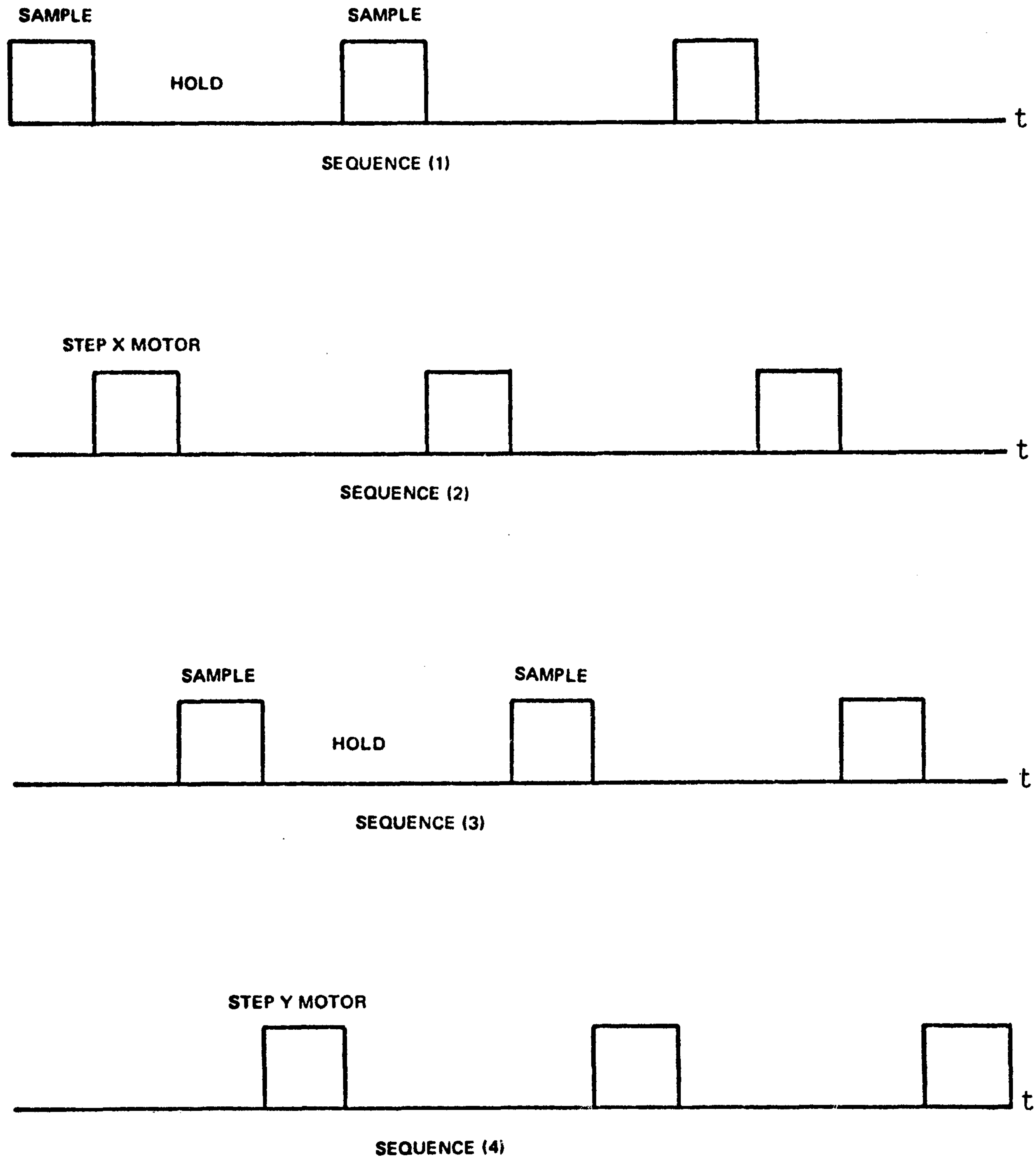

Figure 3-26 STEP-TRACK TIMING SEQUENCES

Thus, if the output of the comparator is a positive voltage after a pulse in the first sequence is applied, the previous Y-axis movement is considered to have been away from the satellite. If the voltage is negative (or zero depending on the type of comparator used), the Y-axis movement is considered to have been toward the satellite. Similarly, if the output of the comparator is positive after a pulse in the third sequence was applied, the previous X-axis movement was toward the satellite, and if the output is negative, the movement was away from the satellite. The logic following the comparator takes this information and generates the appropriate command signals to the stepper motors.

With this type of operation, some average pointing error would obviously exist even if the system is operating under perfect conditions. This is because the antenna's boresight axis is continuously being stepped. The magnitude of the average pointing error depends on a number of system parameters. These parameters are the step-size $\Delta\theta$, the predetection bandwidth B_{IF} and signal-to-noise ratio (in the access control signal demodulator), and the pre-comparator low-pass filter (LPF) bandwidth b.

3.6.2 Performance of the Step-Track Scheme

An analysis of the Step-Track scheme requires the (1) characterization of the antenna gain function (e.g., Equation 3-3), (2) determination of the probability of error when comparing the two sequenced output voltage samples in a noisy environment, (3) derivation of the transitional probabilities between steps and the stationary probability of a given antenna pointing offset, and (4) calculation of the corresponding antenna gain loss as a function of angular step size $\Delta\theta$, detector input S/N, and precomparator (LPF) bandwidth b, for a given set of angular dynamics in relative motion between the satellite and terminal.[38] The critical

(38) Tom, N.N. & Heckert, G.P., Op. Cit.

parameter is the pre-comparator low-pass bandwidth b, which must accommodate the high angular rates inherent with the user ships. Appendix D contains an analysis of the performance of the Step-Track scheme, for sample antenna configurations. The basic example is a fan beam produced by an antenna which is steered in azimuth only. The elevation beamwidth is large enough to cover the expected rolls, etc. (e.g., >±45°) and the azimuth beamwidth is assumed to be 20°. This corresponds to an antenna gain on about 10 dB. It is assumed that the maximum azimuthal angular rate is $\dot{\Theta}$ = 10 deg/sec.

The second case is for a symmetrical beam of width 40 deg (a gain of about 12 dB), with an angular rate of 20 deg/sec in each axis. In each case, a predetection carrier-to-noise density of 40 dB-Hz is assumed, as representative of minumum requirements for the access control modem.

The following Step-Track parameters were concluded as optimum:

$$\text{Step size} \qquad \Delta\Theta = \frac{\Theta_{hp}}{10} \qquad (= 2°)$$

$$\text{Post-Det. LPF Bandwidth} \quad b = \frac{1.5}{T} \qquad (= 15 \text{ Hz}) \tag{3-39}$$

$$\text{Step Period} \qquad T = \frac{\Delta\Theta}{2\dot{\Theta}} \qquad (= 0.1\ °/\text{sec})$$

The stepping period T is the time interval between stepping and sampling, Thus, the time between successive stepping of the antenna beam in the same axis is 4T for dual axis steering and 2T for single-axis steering.

The analysis of Appendix D is not encouraging. Even though the <u>average</u> loss in operational gain due to pointing error was calculated to be only

$$\bar{L} = 1.3 \text{ dB}$$

for the above example, it was that found that the probability of losing system lock is one percent after only 75 steps have been made. With the stepping interval of 0.1 second, this amounts to only 7.5 seconds. The probability of losing lock could be as high as 10% in less than one minute. That is, the above long term average degradation in antenna gain calculated assumes that system lock is maintained. Because of the high angular rate assumed, there would be a high probability of losing lock with the system. If an instantaneous error were such that no signal is received, i.e., the receiver AGC loses lock, then the system would probably not recover and manual re-acquisition would be required. Complete loss of signal can obviously be expected when the satellite is outside the antenna beam, e.g., at angle offsets corresponding to twice the half-power beamwidth. For the example (fan beam) beamwidth (Θ_{hp}) of 20°, the "loss of track" error angle can be taken as ±20°.

The preceding results for the fan beam example are also generally applicable to certain other antennas, when attendant conditions are selected appropriately. Specifically, the results are applicable to a symmetrical antenna beam with the same gain, i.e., an antenna with a beamwidth of about 40 degrees in each axis. This antenna would have to be stepped twice as fast because of the requirement for two-axis stepping. The wider beam and higher rate tend to compensate for each other in terms of performance analysis.

The error calculated, and the probability of loss of lock, could be reduced significantly for higher input carrier-to-noise densities, or for much lower angular rates. While it is possible that the access control channel include multiplexed (working) data as well, such that a total bit rate of 2400 bps and a requirement for a received C/N_0 of 45 dB-Hz exists, the angular rate requirements can not be relaxed. As described in Section 2.4, the angular rates can be considerably higher than 10 deg/sec, especially for the smaller ships.

Another very important factor is the implied rate requirements for the stepping drive of the antenna. In either the fan beam or symmetrical beam antenna examples, the required step period is 0.1 second, in terms of dwell time (per axis). The actual step through $\Delta\Theta$ should be accomplished in at least one-tenth this time i.e., 0.01 seconds. With a two-degree step size, this implies an average drive speed of 200 deg/sec. Clearly, the associated accelerations and peak speeds required to meet the preceding average speed in 0.01 seconds would be a major imposition on an otherwise simple system concept.

Thus, even with higher received carrier-to-noise densities potentially available, it would appear that the Step-Track technique is not advisable for shipboard antennas to be used for satellite communications by the civilian maritime industry.

3.7 COMPARISON OF ANTENNA POINTING TECHNIQUES

The various concepts described in the preceding sections for directing narrow antenna beams at the satellite each have certain merits and disadvantages. Actual selection of a single technique as clearly optimum is not justified at this point, because certain schemes are more suited to certain types of antenna designs. The various approaches to antenna radiator design are discussed in Sections 4, 5 and 6. Following that, a comparison of selected antenna subsystem approaches, including the pointing technique, is presented.

However, certain general conclusions can be drawn:

1. For very low-gain requirements (≤6 dB) only, manual pointing of a movable antenna or manual selection of a small set of switchable radiators is practical, and probably acceptable for low-cost user terminals.

2. For moderate gains of about 7 dB to 14 dB, either slaving to ship's inertial references or automatic tracking by monoscan (single channel monopulse) are most suitable. Selection between these two alternatives should be made upon consideration of specific user vessels for the effectiveness of the additional costs (if any) probable with monoscan, in terms of its negating of the frequent (e.g., daily) realignment and calibration function inherent in slaving.

3. For the higher gains of 15 dB to 18 dB, automatic tracking should be implemented. Of the possibilities in this category, single-channel monopulse is most recommendable because of its relative simplicity.

Table 3-3 summarizes the general comparison of the various antenna beam pointing techniques considered in Section 3. As indicated, the most suitable approach depends on the actual antenna gain considered. Since the baseline range of gain considered in this study is about 10-12 dB, it is of value to examine the cost trade-offs between approaches based on slaving to shipboard inertial references and automatic signal tracking by monoscan. Table 3-4 presents the corresponding comparison.

This latter table presents relative costs only, in that cost of the basic antenna radiator, the transmitter, receiver and related communications equipment is not considered. Nor is the total installation cost considered; the table includes installation costs <u>only</u> where peculiar to the antenna beam positioning subsystem. Clearly, these costs would vary greatly according to the basic antenna concept, i.e., switching of fixed antenna beams, electronically scanned arrays or single beams which are mechanically pointed. This table assumes the latter category as an appropriate reference.

Given that an antenna is remotely steerable by any means, it effectively must have the basic capability of being manually pointed, according to angular positions displayed on a control monitor, e.g., an antenna control unit chassis rack-mounted with the communications equipment in the radio room. The cost of this basic feature is estimated at $2800. (This includes only a portion of the antenna subsystem installation costs, i.e., that associated with the pointing mechanism).

The cost of the equipment peculiar to stabilizing the antenna by slaving to an (existing) gyrocompass and a specially developed low-cost inclinometer (vertical reference) is estimated to be about $1140. When compared to the monoscan approach, the additional installation costs associated with the inclinometer and gyrocompass control lines, etc., must

TABLE 3-3

SUMMARY COMPARISON

OF BEAM POINTING TECHNIQUES

Pointing Technique	Low-Gain Antennas (≤6 dB)	Moderate Gain Antennas (7-14 dB)	High Gain Antennas (15-18 dB)
MANUAL (Operator directed or selected Beams)	Adequate, Least Expensive	Inadequate due to ships Motion	Inadequate due to ships Motion
SLAVING (to shipboard Inertial References)	Adequate, possible gyro-slave only with Low cost	Good Solution, moderate cost requires periodic alignment	Poor Solution
MONOSCAN (Automatic Angle Tracking)	Not cost effective	Good Solution, slightly higher cost	Good Solution, slightly higher cost
STEP-TRACK Automatic	Poor Solution	Inadequate due to high rates, unreliable	Inadequate due to high rates, unreliable

TABLE 3-4

COST COMPARISON OF ANTENNA BEAM POINTING BY SLAVING TO SHIPBOARD REFERENCES AND AUTOMATIC (MONOSCAN) ANGLE TRACKING*

EQUIPMENT BREAKDOWN	SLAVING APPROACH	MONOSCAN APPROACH
1. Basic (Manual) Pointable Mechanism:		
Antenna Pedestal, Gearing	$ 500	$ 500
Drive Motors	400	400
Control Lines, Installation	700	700
Position Display, Manual Controller	900	900
Miscellaneous Furnishings	300	300
Total Common Equipment	$2800	$2800
2. Slaving Equipment, Dual Axis:		
Gyro Repeater (Bearings, Gimbal Ring)	$ 340	
Basic Inclinometer Cost	500	
Inclinometer Repeater (Bearings, Gimbal Ring)	300	
Inclinometer Installation	200	
Installation of Additional Cabling, Gyro & Inclin.	860	
Total Slaving Peculiar	$2200	
3. Monoscan Equipment:		
Monopulse Feed & Comparator		$ 950
Monoscan Converter & Coupler		850
Scan Generator		100
Sync. Angle Detectors, Filters		200
Servo Position Loop Amplifiers		400
Total Monoscan Peculiar		$2500
4. Total Antenna Pointing Subsystem Costs	$5000	$5300

* These costs assume manufacture of approximately 200 units.

be included. These slaving-peculiar installation costs are estimated to be \$1060 for both axes (see Section 3.4.2.3 for vertical reference costs).

The costs of the additional equipment required for monoscan, given that there exists the antenna radiator and basic communications receiver (see Figure 3-18), is estimated to be \$2500. Thus, the total cost of automatic tracking using monoscan is estimated to be \$5300 compared to \$5000 for slaving. Thus, while the monoscan tracking receiver equipment is more expensive than the slaving equipment, the additional installation costs associated with slaved stabilization brings the trade-off to only a few hundred dollars. (No installation costs are peculiar to the monoscan technique; it is essentially self-contained with the antenna and communications equipment respectively.)

This few hundred dollars would appear to be a small price for a ship owner to pay for the deletion of a daily manual alignment problem. While the equipment used in the slaving approach is fundamentally no less reliable than that used for monoscan, the dependence on a crew member for periodic realignment would tend to lower the confidence of a shore based fleet owner in system operation. It might be considered analogous to a hypothetical worldwide telephone system which each individual user is required to "wind-up" his telephone hand set every day or two to stay linked to the system. Would a caller be confident that no answer meant the party was not at home, or would he suspect someone had forgot to wind the telephone?

The costs listed in Table 3-4 are only estimates based on volume (200 or more units) production, derived from experience and a sampling of appropriate equipment suppliers. However, no equipment manufacturer has yet to actually produce analogous satcom terminal equipment in any quantity which is extrapolatable to the subject application.

Thus, the cost trade-offs must be considered in this light. It is anticipated that once a maritime satellite system is operational, different electronic manufacturers could readily market such alternate approaches, with varied quality and reliability. For example, certain ships already equipped with vertical gyros as well as gyrocompasses, and a system of repeaters, may readily elect to employ the slaved stabilization approach, and establish the required periodic realignment procedures.

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SECTION 4

SINGLE BEAM ANTENNAS - MECHANICALLY POINTED

4.1 CATEGORICAL CONSIDERATIONS

4.1.1 General

As discussed in Section 2.2.1, it is considered appropriate in this study to categorize antenna system concepts according to their implied beam-pointing scheme rather than by gain or actual antenna radiator design. Mechanically pointed, "single-beam" antennas are defined herein as those antenna subsystems in which a single radiating element or self-contained assembly of elements is directed in angular position by the remote control of motors. The other two antenna system concepts are switched-beam antennas, in which multiple antennas or radiators are mechanically and electrically fixed at appropriate locations and orientations yielding, together, greater than hemispherical coverage, with beams selected by switching only; and electronically-scanned beam antennas, in which an array of elements is configured to produce a single beam which is electronically steered by continuous control of precision phase-shifters. These latter two categories are considered in Sections 5 and 6, respectively.

The mechanically pointed, single beam category does include planar and linear arrays insofar as such arrays can be used to generate a "narrow" beam, but in this case the whole array would be mechanically pointed. In general, the antennas to be considered here for suitability to the maritime satellite user application include:

- Conventional Parabolic Reflectors
- Parabolic Cylinder with Fan Beam

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- Planar Array
- Linear Array
- Crossed Yagi/Uda Array
- Helical Radiators
- Log-Periodic Antennas
- Conical and Rectangular Horns
- Short Backfire Antenna
- Turnstile on Ground Plane
- Conical Log Spiral
- Cavity-Backed Spirals
- Cavity-Backed Crossed Slots
- Cavity-Backed Crossed Dipoles

As indicated in Section 2.2.2, the important factors to be considered relative to the suitability of the various antenna radiators in this category for the maritime satcom application are

- Effectiveness in range of gain 3 dB to 18 dB
- Amenability to circular polarization
- Practical gain or beamwidth limitations
- Swept volume, overall size
- Amenability to developing a pattern null to facilitate automatic tracking
- Suitability for fan beam applications
- Sidelobe and backlobe characteristics
- Amenability to locating low noise preamplifier and transmitter power amplifier at feed port
- Suitability to radome protection
- Rotary joint and/or cable wrap limitations
- Wind loading and vibration effects

- Relative cost of development and manufacture
- Pecularities relative to installation, if any

Many of these factors are obvious for specific radiator types; others are not so obvious. The design problems associated with each of the antennas considered are reviewed in subsequent sections for the purpose of indicating relative complexity and thus cost. Two of the above factors which are peculiar to t mechanically-pointed, single beam antenna category are the motor drives and the need for rotary joints or cable wrap to transport signals across axes of rotation. These inherent characteristics are frequently considered as two significant disadvantages of the mechanically-pointed antenna category, and are reviewed in the context of the subject application in the following paragraphs.

4.1.2 Antenna Drive and Position Control Considerations

Common to all mechanically-pointed single beam antenna concepts is the use of motors to drive the antenna to desired positions against inertia and wind-torques. Also, angular position data pick-offs at the antenna and corresponding readouts on an antenna control unit housed with the radio equipment are required. The drive motors and gearing should be simple, reliable and operable over long periods without periodic maintenance.

A major factor permitting such simplicity to be realized is that the antenna gains sought are small, which, at L-band, permits use of small antennas and relatively large beamwidths. This is especially true for gains less than 14 dB or so. (This is one major factor leading to the selection of 10 dB as a baseline value of gain in this and related studies.)[3]

For example, with reference half-power antenna beamwidths on the order of 55 degrees to 33 degrees (10 dB to 14 dB gain; see Figure 2-4), the allowable resolution of control devices could be as poor as ±5 degrees. However, conservative design practice dictates a control panel angular position meter resolution of ±1 degree. The angle data pick-offs and controllers with this capability are readily available at low cost.

For the antenna drive motor and control systems, three general categories have been considered as candidates. These are:

1. Open-loop drives, i.e., stepper motors.
2. Closed loop drives, using synchro components as data pick-offs and dc motors (dc torque drive).
3. Closed loop drives, using synchro components as data pick-offs, and ac servo motor drives.

Of these basic categories, which on their own are each acceptable, the third, consisting of synchro components as data pick offs and associated ac servo motor drive, is concluded here to offer the highest reliability and longest life.

For an antenna with a maximum dimension of about 40 cm (about 16 inches) facing the wind, the drive torques required should be no more than about 20 to 40 meter-newtons (15 to 30 ft-lbs) in wind speeds of 145-200 km/hr (75 to 110 knots). This maximum dimension is typical of antenna candidates in the gain range 10-15 dB.

The torque requirement of the servo drive for a 40 meter-newton maximum wind load would normally call for a dc torque drive since the choice of an ac servo drive is generally confined to "smaller" torque requirements. Manufacturers are, however, available for ac servo

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(3) 1970 DOT-CG Study, Op. Cit. Volume III.

motors to satisfy these requirements, although not as numerous as the "instrument" servo motor suppliers. The ac servo motor would fall into a 1/2 or 3/4 HP category (e.g., $\leq$10 cm or 4 inches diameter). The associated gear train would fall into a larger size class of gears normally associated with instrument servos. Typically the gear face width should be less than 1.3 cm.

As implied above a dc drive has basic advantages and limitations relative to an ac drive; and the final choice of an ac drive system is based on the particular civilian shipboard user application being considered. Some of the pertinent trade-offs considered were:

1. The dc motors can deliver a high output from a small frame size, since there are no slip losses as in an ac motor.
2. The dc motor circuitry permits use of a variety of stabilization techniques.
3. Residual voltages are usually negligible in dc motor circuits.
4. DC motors are more efficient than ac motors, especially in applications requiring variable speed.
5. It is possible to employ dc motors in a manner in which no power is required under no-signal conditions. (In contrast, ac motors require main field excitation.)

The aforementioned advantages of a dc motor over an ac servo motor are considered offset by the following disadvantages for this application:

1. DC motors have commutators that require periodic maintenance checks to ensure reliable brush performance. Such maintenance is a serious problem area in complex or remote

assemblies, such as atop a mast. Brushes represent the greatest single deterrent to a more widespread use of dc motors.

2. Radio interference generated by brushes often leads to special filtering and shielding requirements to reduce this interference to a satisfactory level.
3. The use of induction-type data pick-offs (as would be used in an ac approach), which includes the category of synchros and resolvers, avoids the requirement for demodulators and dc circuitry associated with the use of dc motors.
4. High field strengths used in dc machines increases cogging and therefore decreases small signal sensitivity. This is further accentuated by brush friction.

The above discussion is considered to be a summary of the trade-offs pertinent to ac vs dc closed-loop drives. The open loop drive, listed as the first category, has most of the disadvantages associated with the maintenance of dc brushes, along with the other hardware limitations of friction-type surfaces. The single most dominant factor in a choice of an ac closed loop drive is that all the rotating components, namely, motor and data pick-offs, are essentially frictionless (with the minor exception of slip rings associated with the angle data pick-offs).

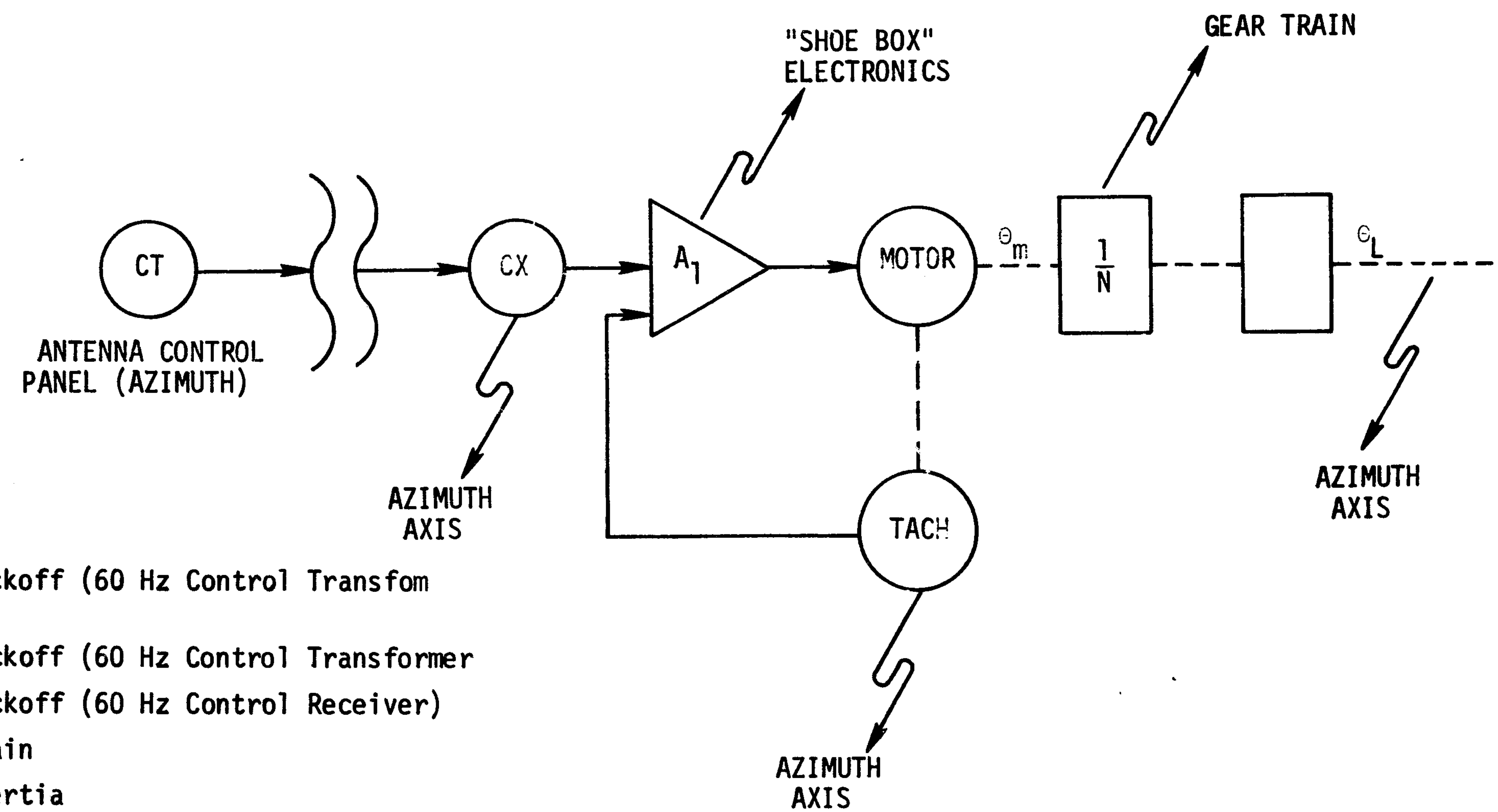
With regards to differences between an azimuth-only drive motor for single-axis drive and the use of both azimuth and elevation drive motors for dual-axis operation, the servo loops are essentially the same with the following exception. The two-axis drive would contain a variable gain amplifier in the azimuth axis whose gain would be a function of the elevation angle. This "secant function" sensor is readily obtained from the same data pick-off (resolver or synchro type), which is used in the elevation axis shaft. The elevation axis sensor would provide the necessary gain change information feeding the azimuth loop servo. This function is required to compensate for the decrease in azimuth angle changes, when referred to the horizon plane, as elevation angle increases, as discussed in Section 2 and 3 previously.

Figure 4-1 shows a basic configurational diagram of a motor drive system for a single axis (azimuth). Figure 4-2 shows the typical servo block diagram, indicating the transfer functions, position feedback, etc. of the elevation axis. Figure 4-3 shows the equivalent block diagram for an azimuth axis, which includes the secant correction function. The control input to the positioning system may be a manual command from a control panel, or a gyro-repeater output, or the dc output of an automatic angle tracking receiver.

Of course, the use of azimuth and elevation is, at this point, an example. Other mount arrangements can be used, such as an X-Y mount which consists of two orthogonal elevation axes. In any case, the conclusion herein is that both the torque and accuracy requirements are very modest, and that an ac drive system is most suitable from a reliability standpoint. It is estimated that such a dual-axis servo drive system, including the angle data pick-offs, meters, motors, gears, etc. would cost less than $2000, when assembled with generally off-the-shelf components from existing manufacturers. Dedicated manufacture in volume would reduce these costs.

4.1.3 <u>Rotary Joints and Cable Wraps</u>

One particularly undesirable feature associated with mechanically pointed antennas is the need for a special rotatable transmission line to pass the transmit and receive signals through the axes intersections of the antenna pedestal. In general, there are three basic techniques for implementing such a requirement. These are rotary joints, cable wraps and slip rings.



C_T = Data Pickoff (60 Hz Control Transfom

C_T = Data Pickoff (60 Hz Control Transformer

C_X = Data Pickoff (60 Hz Control Receiver)

N = Gear Train

J_L = Load Inertia

Θ_m = Motor Angle

Θ_L = Load Angle

Figure 4-1 BASIC CONFIGURATION OF AC SERVO MOTOR DRIVE SYSTEM FOR ONE AXIS (AZIMUTH)

K_1 = Data Pickoff Gradient
1 volt/degree (typical)

K_T = Motor Torque Constant: ft-lbs/volt

K_o = Tachometer Gradient: volts/RPM

J = Load Inertia

D = Internal Motor Damping

Θ_e = The angular difference between the control input and antenna axis

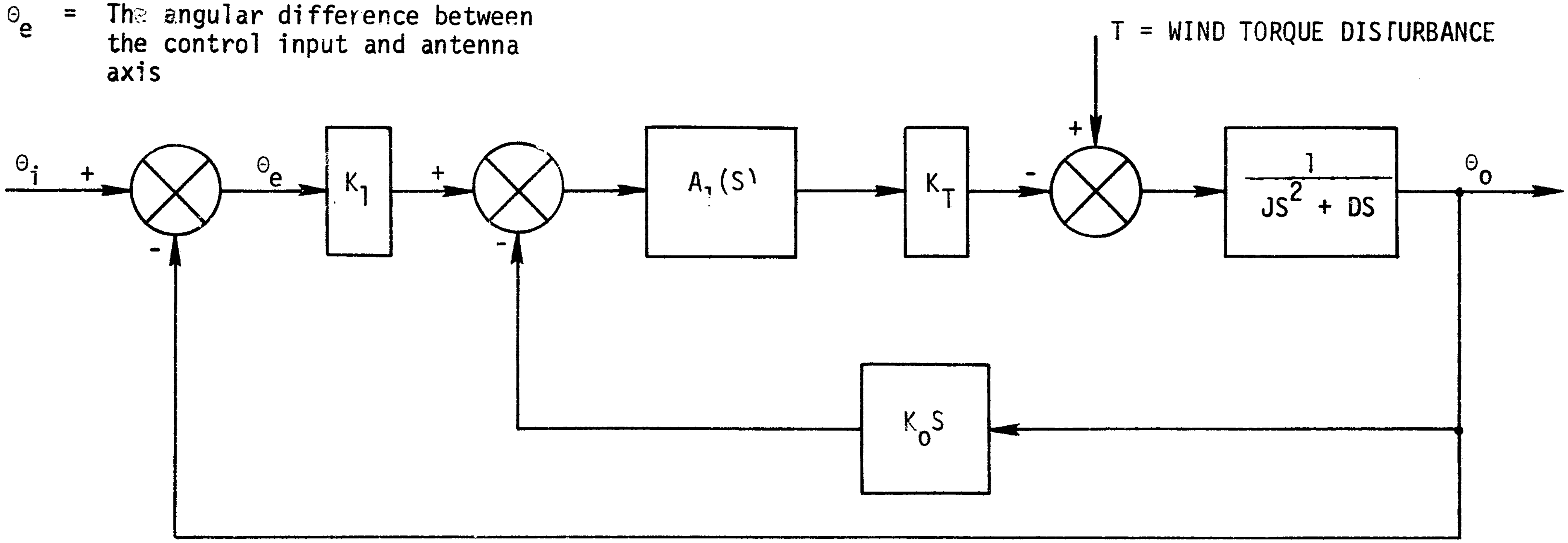


Figure 4-2 FUNCTIONAL BLOCK DIAGRAM OF ELEVATION AXIS SERVO DRIVE

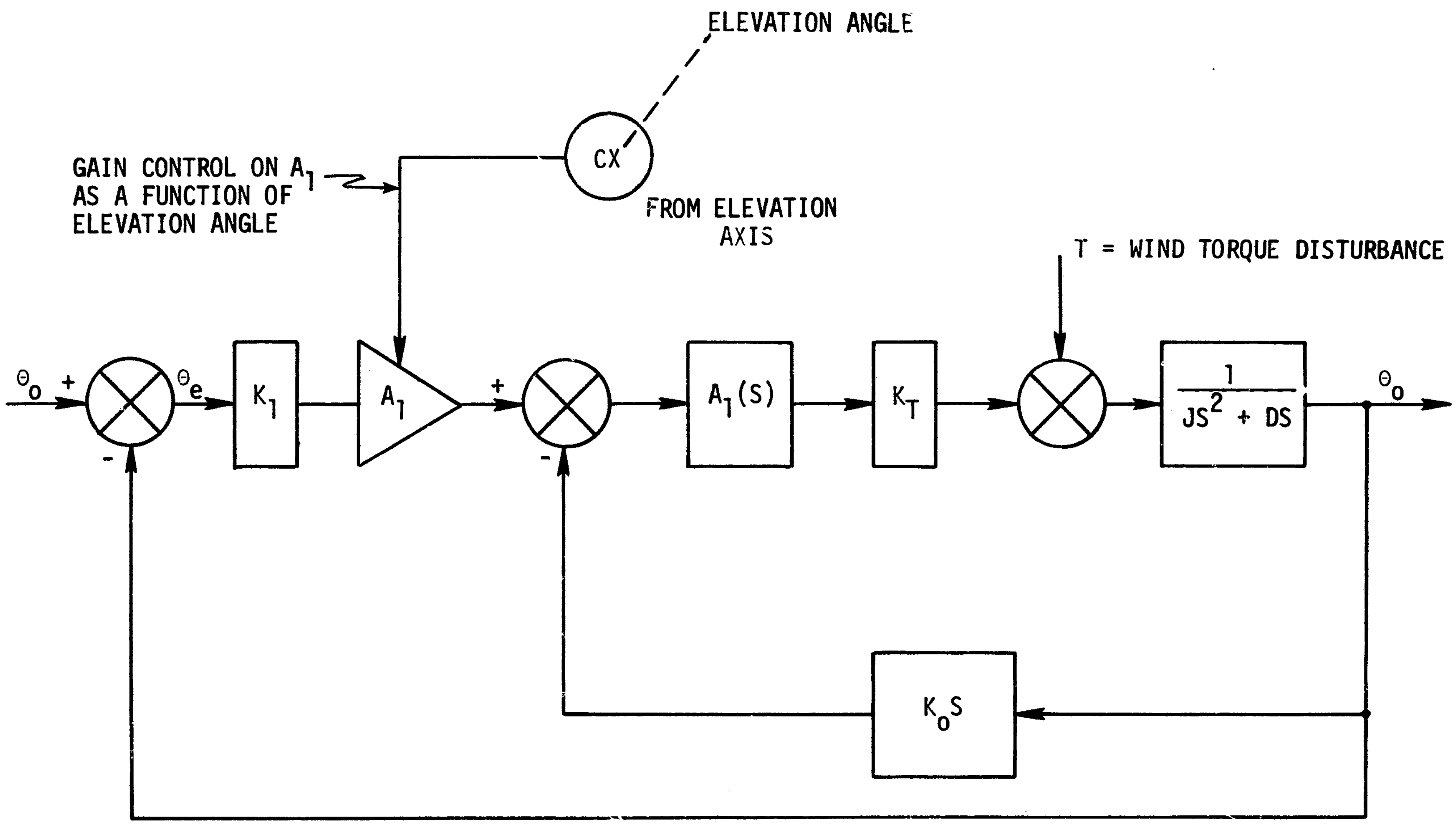


Figure 4-3 FUNCTIONAL BLOCK DIAGRAM OF AZIMUTH AXIS SERVO DRIVE

In typical microwave antenna systems where very low line loss is desired, rotary joints are used, with waveguide interface. While rotary joints can be very reliable, they are typically expensive relative to the low-cost shipboard terminals being considered here. They are also limited to fairly narrow-band applications. However, this does not present any difficulty in the maritime satellite application. In quantity production, of course, their costs are by no means prohibitive. Virtually all large ships are equipped with X-band (9 GHz) radars which are continually scanned in azimuth, (i.e., a mechanically-pointed fan beam) with waveguide running up the mast to a horn antenna with a rotary joint transition. The cost of a complete radar set, including antenna, drive, transmitter, receiver and display costs, is in the range of $12,000 to $18,000 depending on manufacturer and quality.

The most desirable transistion would be a cable wrap. The somewhat higher line loss associated with cable at L-band is readily permissible in the subject application. The low noise preamplifier would be sufficiently small, so that it can, in fact, be located adjacent to the antenna prior to axes transitions in the receive path. While the low-power transmitter should also be located at the antenna, it may not be practical to actually place it "above" the two axes. In this case, however, the small cable loss would be insignificant. The only constraint that would negate the use of cable wrap exists in the azimuth axis of an elevation-over-azimuth gimbal system, i.e., mount. A cable wrap can readily satisfy the elevation angle excursion, e.g., from -45° to 135° in the worst (small ship) case. But some users may define a need for operation while continuously turning through many circles. A cable wrap could satisfy a certain amount of turning, e.g., as much as 720° in azimuth.

For many applications, this should be ample. For the vast majority of commercial ocean-going vessels which by and large travel from A to B, a cable wrap should suffice even in azimuth. If the cable eventually reaches a limit, the set operator would have to go "off the air" for a minute or so while he (remotely) rotated the antenna by a full turn or so. Generally, this could be done in port between trips. However, it is possible, of course, to envision certain situations where ships needed to maneuver or simply circle, perhaps in rescue operations.

Two alternatives exist to the requirement for rotary joints in all ships. First, a rotary joint can readily be an optional component in a terminal, to be purchased at the discretion of the individual user. Second, an X-Y antenna mount can be used. With dual elevation axes providing the required hemispherical coverage instead of an elevation-over-azimuth mount, simple cable wraps may be employed.

The disadvantages of an X-Y mount are two-fold. Analogous to the rate limitation at the zenith in az-el mounts, producing a "keyhole" in the coverage, two "keyholes" exist at (opposite ends of) the horizon due to finite rate limitations with X-Y mounts. However, with the subject beamwidths being so wide, actual holes in the coverage would not in practice exist. At those certain directions where rate limitations create a pointing error, the effective gain would be reduced, but system lock should not be lost.

The second disadvantage of X-Y mounts is that of operator control. As noted in Section 3.3, every directable antenna should have a manual control mode, even if automatic tracking or slaved stabilization is also used. It is necessary for initial acquisition, alignment, etc. The display of two elevation axis angles is not convenient for use by the operator, and for this reason an X-Y to Az-El coordinate translator should probably be employed. Such devices have been used previously. They are generally electro-mechanical. This added component is undesirable in that it would be an added cost to the antenna subsystem, but this additional cost need not necessarily be significant.

The third alternative, slip rings, is applicable only at low frequency, e.g., 5 MHz and below. This would not be a suitable frequency for transmission of signals between the antenna location and the radio room, clearly, with spectrums of up to 8.5 MHz in each transmission direction. Thus, the trade-off may be essentially reduced to a cost comparison between a rotary joint and an X-Y to Az-El coordinate translator. Unfortunately, there is little evidence currently which suggests any significant difference between the two alternatives when large quantity manufacture is considered. The mount/axes trade-off is also affected by other considerations. An X-Y mount may actually have advantages in terms of dynamics, in that one axis can be parallel to the roll axis. The actual size and geometry of the antenna radiator may also be a factor. The following sections review these and other characteristics of various single-beam antenna radiators suitable for mechanical pointing.

4.2 CONVENTIONAL PARABOLIC ANTENNAS

4.2.1 General Characteristics

By far, the most used antenna for satellite communication terminals is the conventional parabolic dish. It is highly efficient for very high gain applications, e.g., 20 to 60 dB. The antenna, a paraboloidal surface of revolution, is well known for its ability to focus RF energy as an optical device. Its performance depends on obtaining an efficient combination of reflector and feed geometry. Parabolic antennas can be designed for very low sidelobes and backlobes, and, with specially shaped reflectors which deviate slightly from paraboloids and cassegrain subreflectors, very high efficiencies (e.g., 75%) may be obtained when the size to wavelength ratio is large. The desired circular polarization is also readily achievable.

The reflector size is generally selected on the basis of a gain requirement, while the feed is designed for efficient illumination of the reflector surface. Generally, a near optimum performance is obtained when the reflector edge illumination is approximately 10 dB below that of the center. Higher illumination tapers than this are employed generally for sidelobe performance improvement at the expense of a slight decrease in gain.

An ideal circular aperture antenna is one which is illuminated by a point source with a uniform phase and amplitude distribution over the surface of the dish and no spillover loss. Moreover, the gain of the antenna is directly proportional to the physical aperture, A , i.e., ideally:

$$G = \frac{4\pi}{\lambda^2} A \qquad (4\text{-}1)$$

where λ = signal wavelength

$A = \pi d^2/4$

d = antenna diameter

In actual practice, the gain performance departs appreciably from this ideal value, mainly due to a departure from uniform illumination and other inadvertent losses of energy due to spillover, aperture blockage, cross-polarization and ohmic losses. The gain factor, G, is typically lower by an efficiency factor, η, which has a value between 0.5 to 0.7 in most applications. As noted previously (with Figure 3-1),

$$
\begin{aligned}
G &= \frac{4\pi}{\lambda^2}\left(\frac{\pi d^2}{4}\right)\eta \\
&= \left(\frac{\pi d}{\lambda}\right)^2 \eta \\
&\simeq 5\left(\frac{d}{\lambda}\right)^2
\end{aligned}
\tag{4-2}
$$

Other pertinent relationships between antenna gain, beamwidth and aperture diameter are given below:

$$\Theta_{hp} = \frac{k_i \lambda}{d} \tag{4-3}$$

$$G = 0.5\left(\frac{\pi k_i}{\Theta_{hp}}\right)^2 \tag{4-4}$$

$$G(\Theta) = G - 12\left(\frac{\Theta}{\Theta_{hp}}\right)^2 \quad \text{(in dB)} \tag{4-5}$$

where k_i = 69, in deg. (beamwidth constant for 10 dB illumination taper)

Θ_{hp} = half-power beamwidth in degrees

G = peak gain of the antenna

G (Θ) = gain at angle (Θ) from beam center

The parabolic reflector may be excited by a feed horn whose dimensions may be determined by the empirical formulas given, for example, by Silver:[21]

For the electric plane:

$$\Theta_{10}(E) = \frac{88\lambda}{B} \text{ (degrees)}, \quad \frac{B}{\lambda} < 2.5 \tag{4-6}$$

For the magnetic plane:

$$\Theta_{10}(H) = 31 + 79\frac{\lambda}{A_m}, \quad \frac{A_m}{\lambda} < 3 \tag{4-7}$$

where $\Theta_{10}(E)$ and $\Theta_{10}(H)$ are 10 dB beamwidths

B = aperture in electric plane

A_m = aperture in magnetic plane

4.2.2 Low-Gain Parabolas at L-Band

The typically-used feed horn dimensions at 1600 MHz are generally quite large and thus waveguide feeds for such an antenna should be avoided. A better choice of feed design for an L-band parabolic antenna with a gain of 18 dB or less would be a crossed dipole backed by a metal cup. Shown in Figure 4-4 is a typical design employing cup-dipoles in an orthogonal arrangement.

The diameter, d, is selected in accordance with Equations (4-2) or (4-3) to satisfy the specified gain or beamwidth. Ideally, a 75 cm (30 in) diameter would be required for a gain of about 18 dB. A gain of 15 dB

(21) Silver, S. Op. Cit.

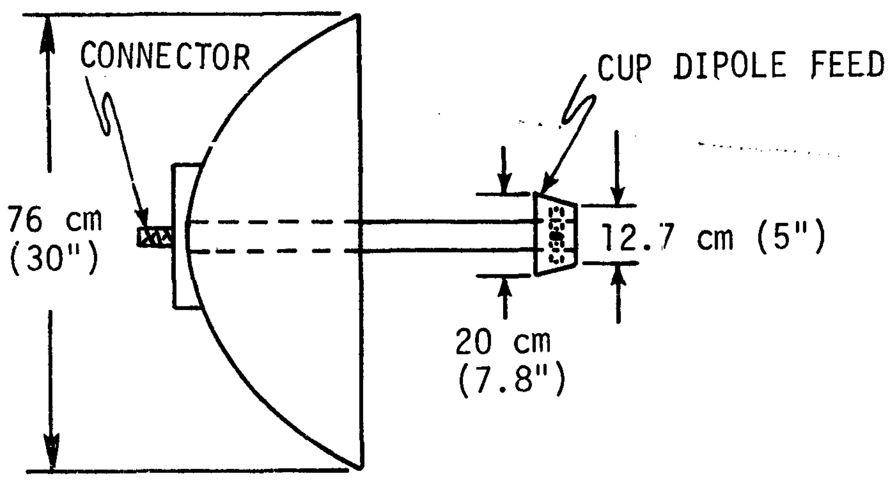


Figure 4-4 TYPICAL PARABOLIC REFLECTOR ANTENNA

requires about a 50 cm (20 in) diameter. (See Figure 3-2.) The size of cup is normally selected to provide proper 10 dB beamwidth for the chosen focal length and aperture dimensions. The subtended angle from the focal point to the edges of the reflector defines the required 10 dB beamwidth. Figure 4-5 presents the subtended angle values for various F/d (focal length to diameter ratio) values.

The feed angle corresponding to the reflector edges may be determined from Figure 4-5. One additional factor due to space attenuation from the focal point to the edge of the reflector must be considered before the 10 dB beamwidth for the feed may be specified. The space attenuation vs feed angle is presented in Figure 4-6.

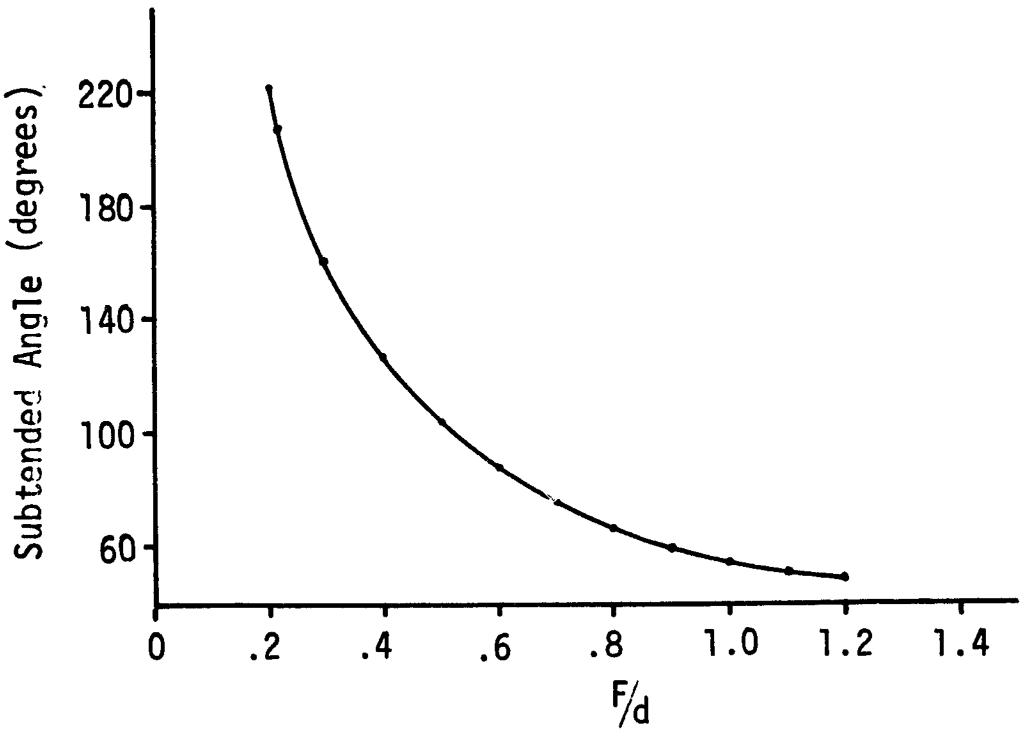


Figure 4-5   RATIO OF FOCAL LENGTH TO APERTURE DIAMETER (F/d) VS SUBTENDED ANGLE AT FOCAL POINT

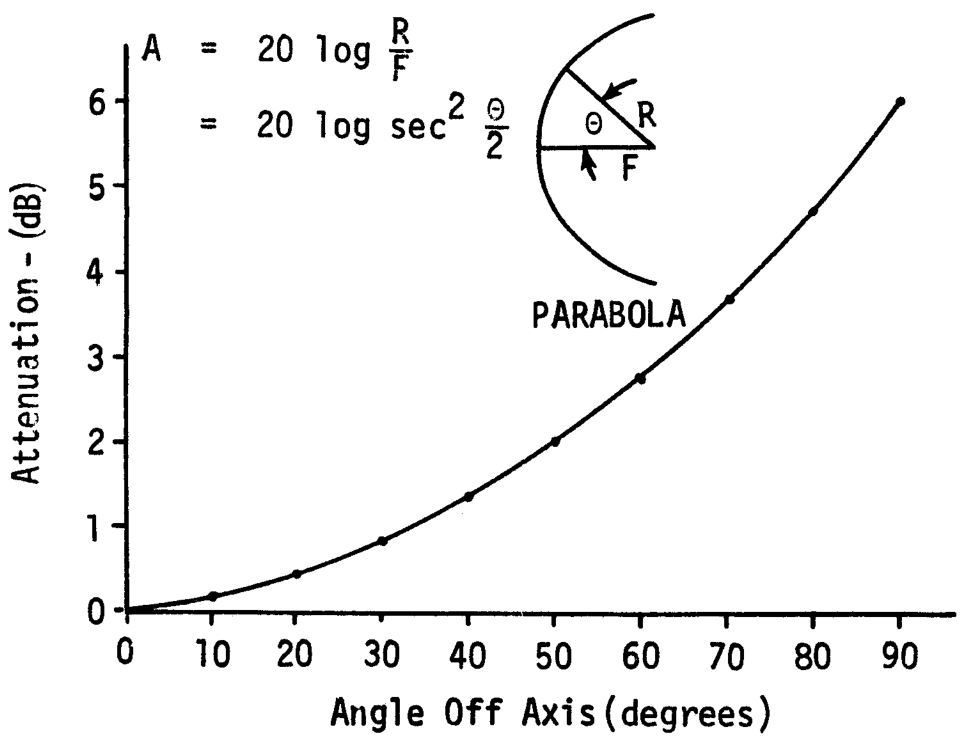


Figure 4-6   SPACE LOSS VS SUBTENDED ANGLE FOR PARABOLOID

As an example of how Figures 4-5 and 4-6 may be utilized in feed geometry selection, a typical F/d ratio of 0.4 can be assumed. From Figure 4-5, the corresponding subtended angle is seen to be 128°. From Figure 4-6, a space attenuation of 3 dB may be noted, which means that a 7 dB beamwidth of 128° is required of the feed to obtain 10 dB illumination taper at the reflector. While this could be achievable, the feed is large and constitutes significant aperture blockage.

In general, the advantages of parabolic antennas are most significant when high values of gain (>20 dB) are specified. At values below this, the antenna diameter is not large relative to the wavelength, and efficiency is poor. This results in a physically large antenna, compared to some other candidates, for the general shipboard terminal requirements of gain and beamwidth. While aperture (size) efficiency is not considered of primary significance for low gain antennas in the subject shipboard application, it would be considered important for values of gain above 14 dB, where diameters on the order of 60 cm to 90 cm (i.e., 2 ft. to 3-1/2 ft.) are implied. This size of antenna, with its attendant swept-volume, tends to be quite significant in terms of a burden on shipboard operation, maintenance and installation. The torques required to drive the antenna in the face of high winds also get large, increasing as the square of diameter.

4.3 PARABOLIC CYLINDER WITH FAN BEAM

The parabolic cylinder is a reflector antenna which can be used to obtain an asymmetrical fan beam with very large aspect ratio (length-to-width) which is not easily obtained from a section of a paraboloid or other antennas. It is practical to use a parabolic cylinder to achieve an aspect ratio greater than 8:1. The beamwidth of the narrow coverage, Θ_{hp}, is mainly controlled by the width of the aperture, d, which can be determined conveniently from Equation (4-3), i.e.,

$$\Theta_{hp} = k_i \left(\frac{\lambda}{d}\right) \tag{4-3}$$

where k_i is a factor of about 70 (in degrees) related to a 10 dB illumination taper.

The broad coverage of the fan beam with the cylindrical reflector excited by a dipole is determined by the pattern of the dipole. In this case the antenna behaves like a corner reflector except that the dipole must be placed at the focal line of the parabolic cylinder. The polarization characteristic of this antenna is linear along the dipole plane. A polarizer must be employed to change the inherently linear polarization characteristic to circular. The polarizer may be located at the reflector aperture as shown in Figure 4-7. This antenna has been dimensioned to provide approximately 11 dB peak gain.

The antenna provides a 17° x 120° fan beam. This configuration is suitable for one dimensional mechanical pointing such as by slaving only to a gyrocompass. It would, however, have high axial ratios near the edge of the wide-beam coverage, and generally be subject to the problems associated with a single-axis pointed fan-beam, as discussed in Sections 2.4 and 3.4.

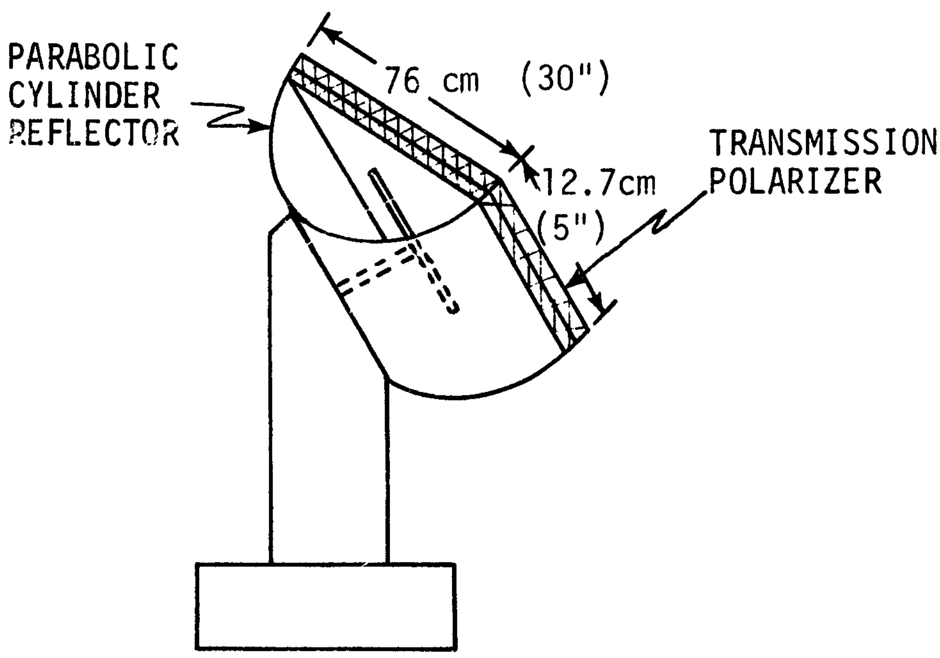


Figure 4-7 PARABOLIC CYLINDER ANTENNA WITH 120° x 17° FAN BEAM

4.4 PLANAR ARRAY

4.4.1 General

A planar array of the most commonly used form is shown in Figure 4-8. The antenna elements are arranged in rows and columns by maintaining a constant separation from one another. The antenna is easily analyzed in terms of linear array theory. The number of elements to be used in each row or column is usually selected in accordance with beam shape and gain required. Consequently, arrays of rectangular or triangular shapes are often used.

4.4.2 Symmetrical Beams

The square aperture array presented in Figure 4-8 is designed mainly to provide a high gain symmetrical beam. The elements in the array are all excited with equal amplitude and phase. The excitation circuit is of corporate feed design consisting of power dividers and equal length transmission line sections. The detail design of the circuit is shown in Figure 4-9. The directivity of the uniform distribution array is given by

$$D = \pi D_x D_y \qquad (4\text{-}8)$$

where D_x and D_y are directivities of the row and column linear array, respectively. Tapered distributions usually give lowered sidelobes at the penalty of some loss in directivity. In terms of design implementation, the uniform excitation is by far the simplest since the feed design is common throughout.

For the array shown in Figure 4-8, D_x and D_y are equal since there are the same number of elements in rows as in columns.

The advantage of a planar array is the great flexibility in obtaining a shaped beam by varying the excitation and grouping of the array elements in a particular manner. In comparison with parabolic reflector designs, the array efficiency is higher for obtaining a non-scanning broadside beam. The cost is higher than for a parabolic dish.

It is estimated that the 4 x 4 array, shown in Figure 4-8, provides a gain of 15.5 dB. The half power beamwidth is approximately 28.5°.

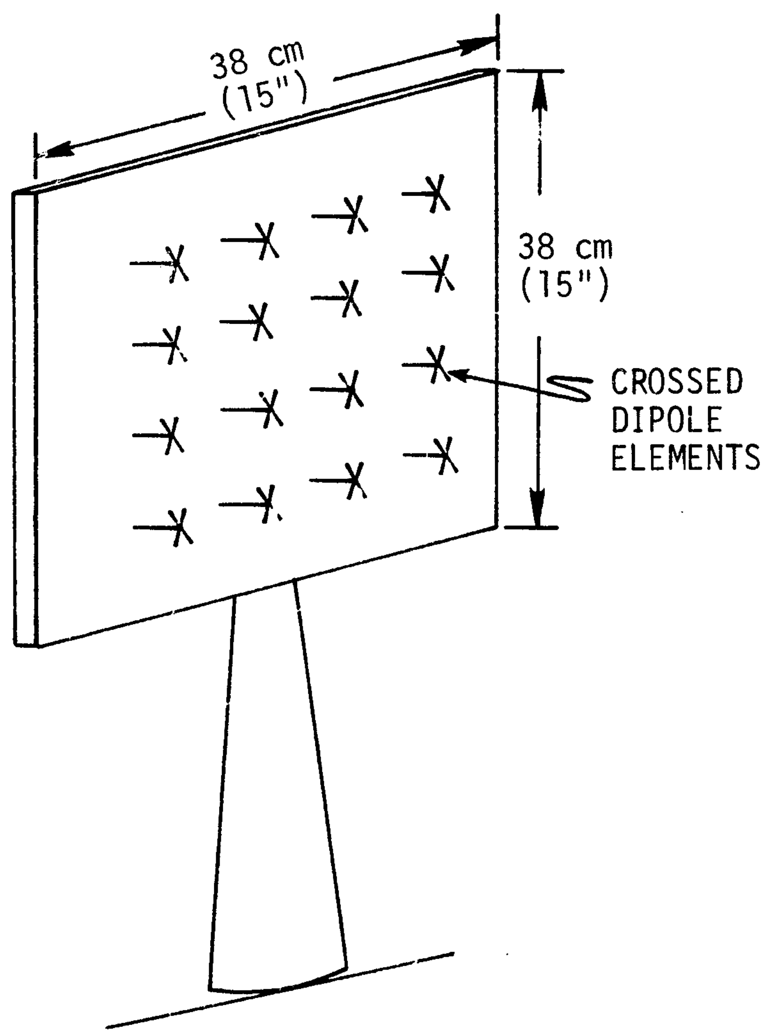


Figure 4-8   4 x 4 CROSSED DIPOLE PLANAR ARRAY

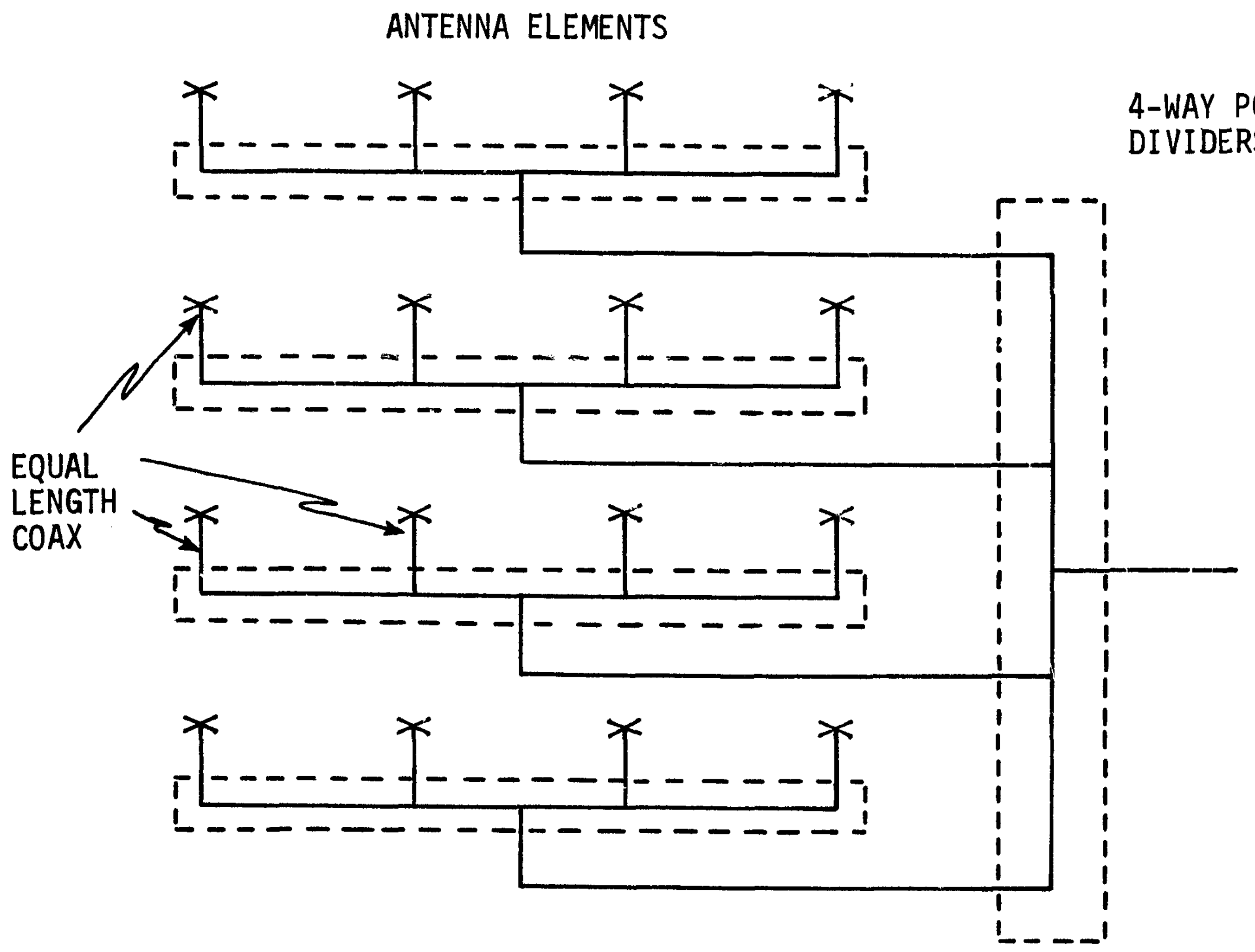


Figure 4-9 FEED ARRANGEMENT FOR PLANAR ARRAY

4.5 LINEAR ARRAY

4.5.1 General

The linear array is an antenna design by which a fan beam radiation pattern of very large aspect ratio can be provided. The elements of the array are arranged in a straight line in one dimension with a constant separation. In a non-scanning design, the elements can be most conveniently excited uniformly in amplitude and phase by means of power dividers and equal line lengths. Under this condition, maximum gain is obtained. The direction of maximum radiation is perpendicular to the array aperture. The narrow beamwidth of the fan beam is primarily determined by the number of elements, the beamwidth decreasing as the number of elements increase. The broad coverage beamwidth of the fan beam is approximately equal to the beamwidth of the individual element, which is, however, typically limited in size to less than that desired. For maximum gain performance, the element spacing is commonly held to 0.7 wavelength. The array beam polarization characteristic is dictated by that of the element.

4.5.2 Circular Polarization and Coverage

A circularly polarized array beam can be obtained by arraying spiral, crossed slots or crossed dipoles or other circularly polarized radiators. The major problem is how to provide low axial ratio circular polarization over a wide angle sectoral beam. One way this can be done is with a linear array of several multi-arm conical spirals (see Section 4.12) The wide angle beam and circular polarization properties of 4 and 8 arm conical log-spirals discussed in Section 4.12 permit fairly low axial ratios over as much as a 180° beamwidth cardioid-of-revolution beam shape, with very low (<-20 dB) back radiation. Thus, a linear array of 4 such elements would be arrayed as illustrated in Figure 4-10.

Although n-element conical spiral antennas are capable of under 3 dB axial ratios over a hemisphere, it may be desirable to hold axial ratios even lower than this figure. To avoid polarization decoupling losses of several tenths of a dB, axial ratios of under 1.5 dB would be required. A novel technique, wherein the polarization ellipse of paired elements in an array are crossed, permits a resultant lowering of array axial ratio below the axial ratios of the array elements themselves. This scheme is referred to as the "crossed-ellipse array" concept. Basically, it depends upon recognition of the fact that the superposition of crossed ellipses of equal axial ratio and intensity can result in a lower axial ratio for the array than that of its elements.

With this concept, the lower the axial ratios of the crossed ellipses, the greater is the tolerance to error in matching the two ellipses properly. Figure 4-11 shows the calculated resultant axial ratio vs the error in amplitude balance of the two waves. Figure 4-12 shows the effect of phase or ellipse orientation error on resultant axial ratio. Figure 4-13 indicates axial ratio unbalance vs resultant axial ratio performance. These plots show that superimposed crossed ellipses, fed in phase, result in a lower axial ratio for given excitation errors than conventional crossed quadrature-phased linear polarized fields.

Figure 4-14 shows the beam area of two crossed fields of quasi-parallel ellipses, illustrating the desired polarization orientation of crossed element pairs. Figure 4-15 contrasts the desirable quasi-parallel element ellipses with undesirable patterns of quasi-radial ellipses which cannot be rotated to produce crossed sets of ellipses, except on axis.

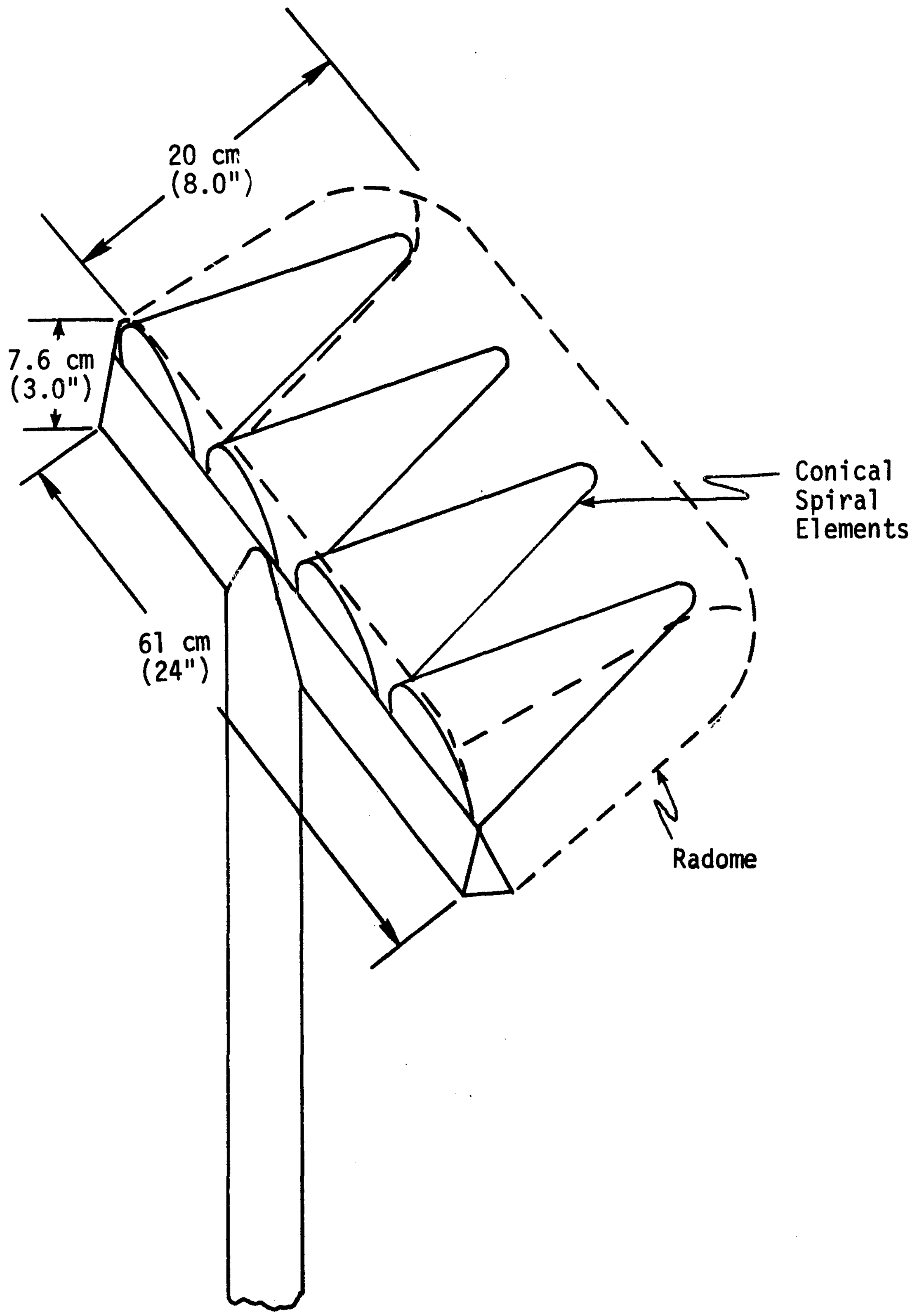


Figure 4-10 FAN BEAM CON-SPIRAL ARRAY

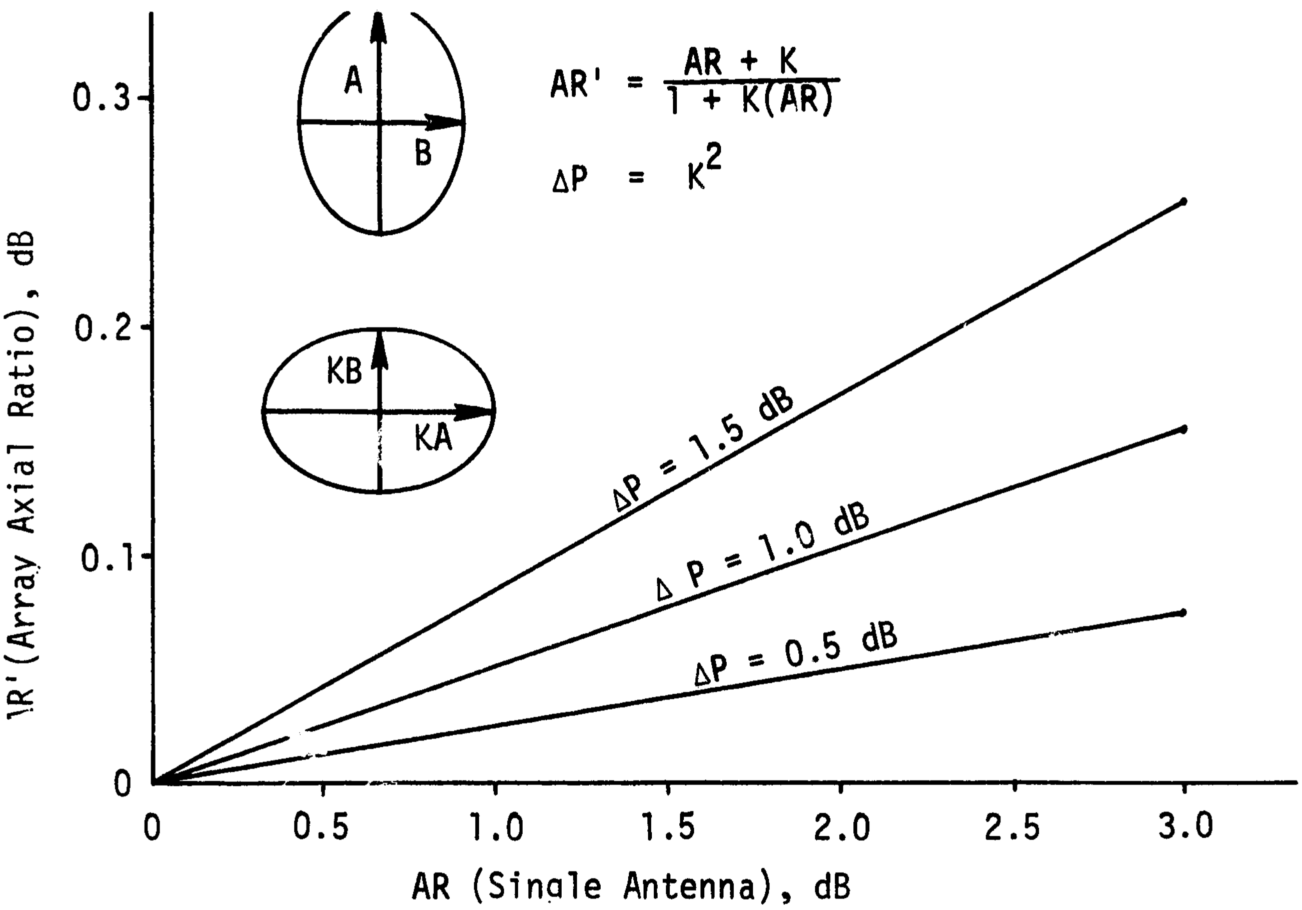


Figure 4-11 EFFECT OF UNEQUAL POWER SPLIT ON AXIAL RATIO (AR) OF A TWO-ELEMENT CROSSED ELLIPSE ARRAY

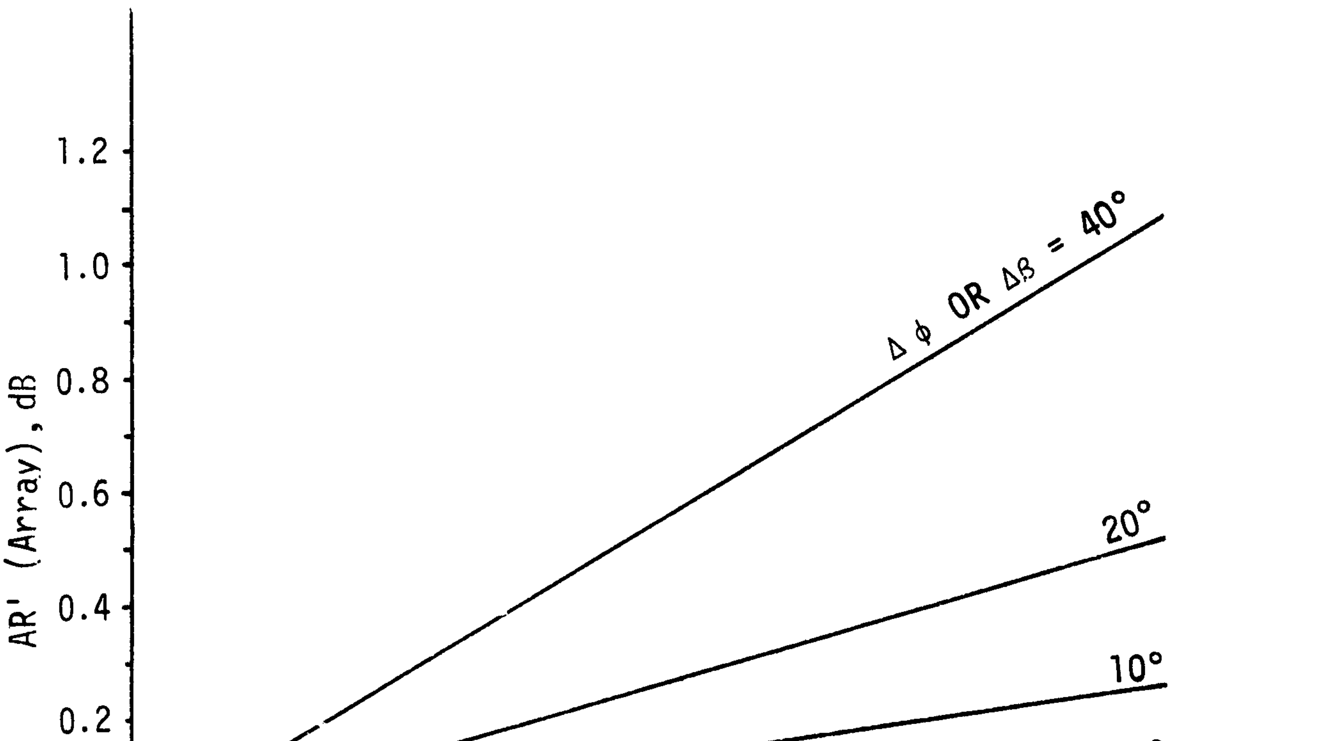


Figure 4-12 EFFECT OF PHASE ERROR (Δφ) OR ELLIPSE ORIENTATION ERROR (Δβ) ON AR IN A TWO-ELEMENT CROSSED ELLIPSE ARRAY

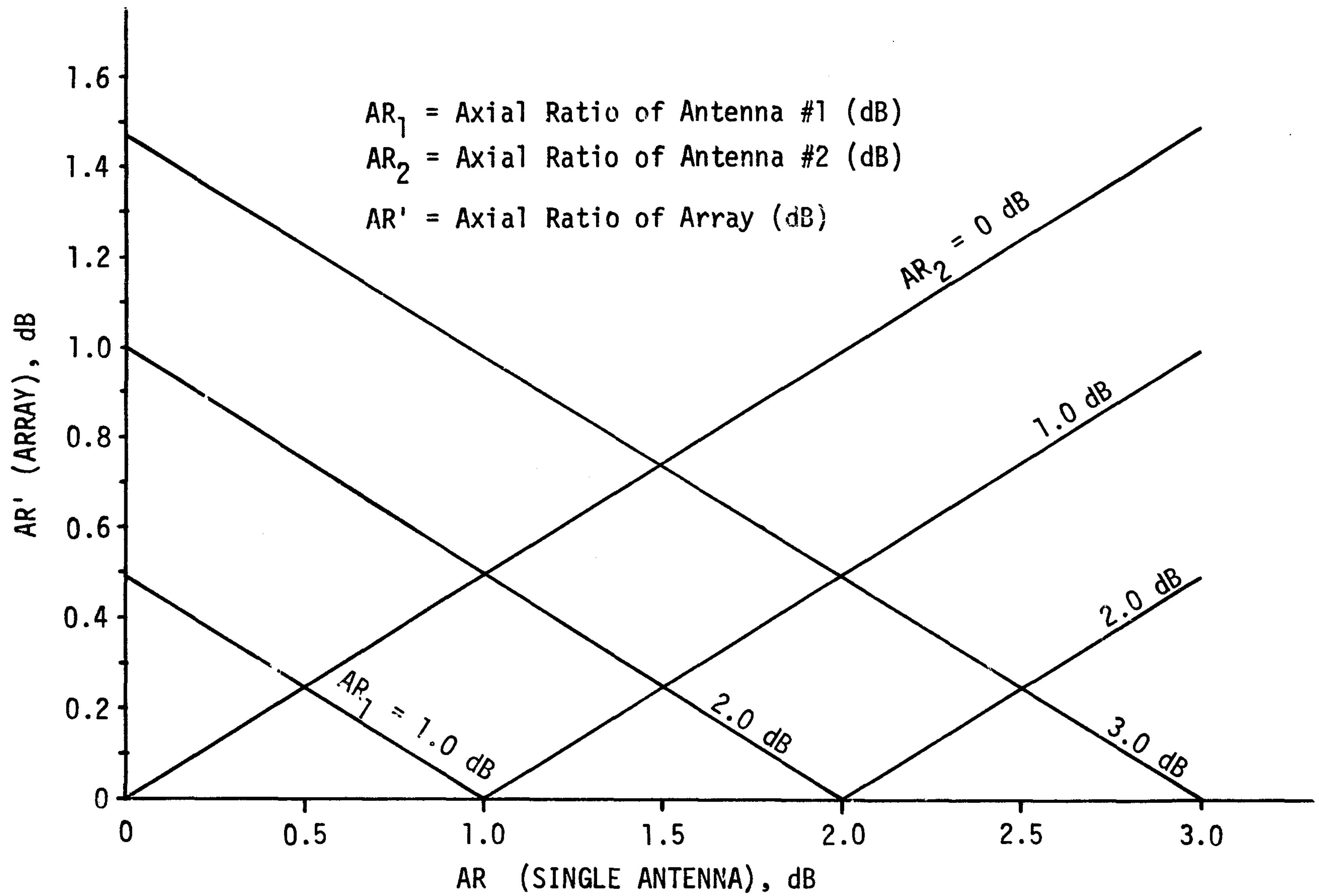
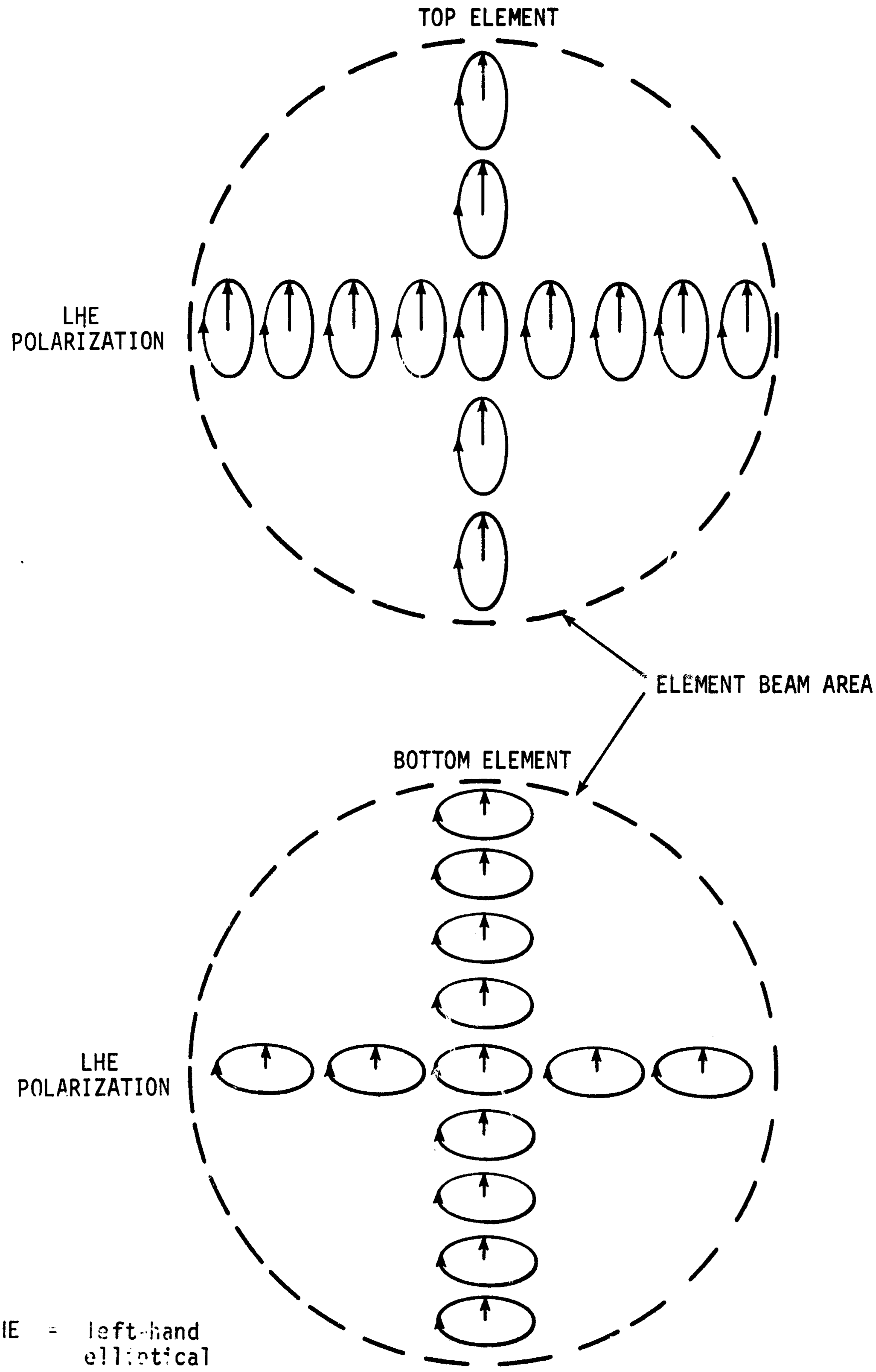


Figure 4-13 EFFECT OF UNEQUAL AXIAL RATIOS ON THE ARRAY AXIAL RATIO OF A TWO ELEMENT CROSSED ELLIPSE ARRAY



LHE = left-hand elliptical

Figure 4-14 CROSSED ELLIPSE CP ARRAY PRINCIPLE

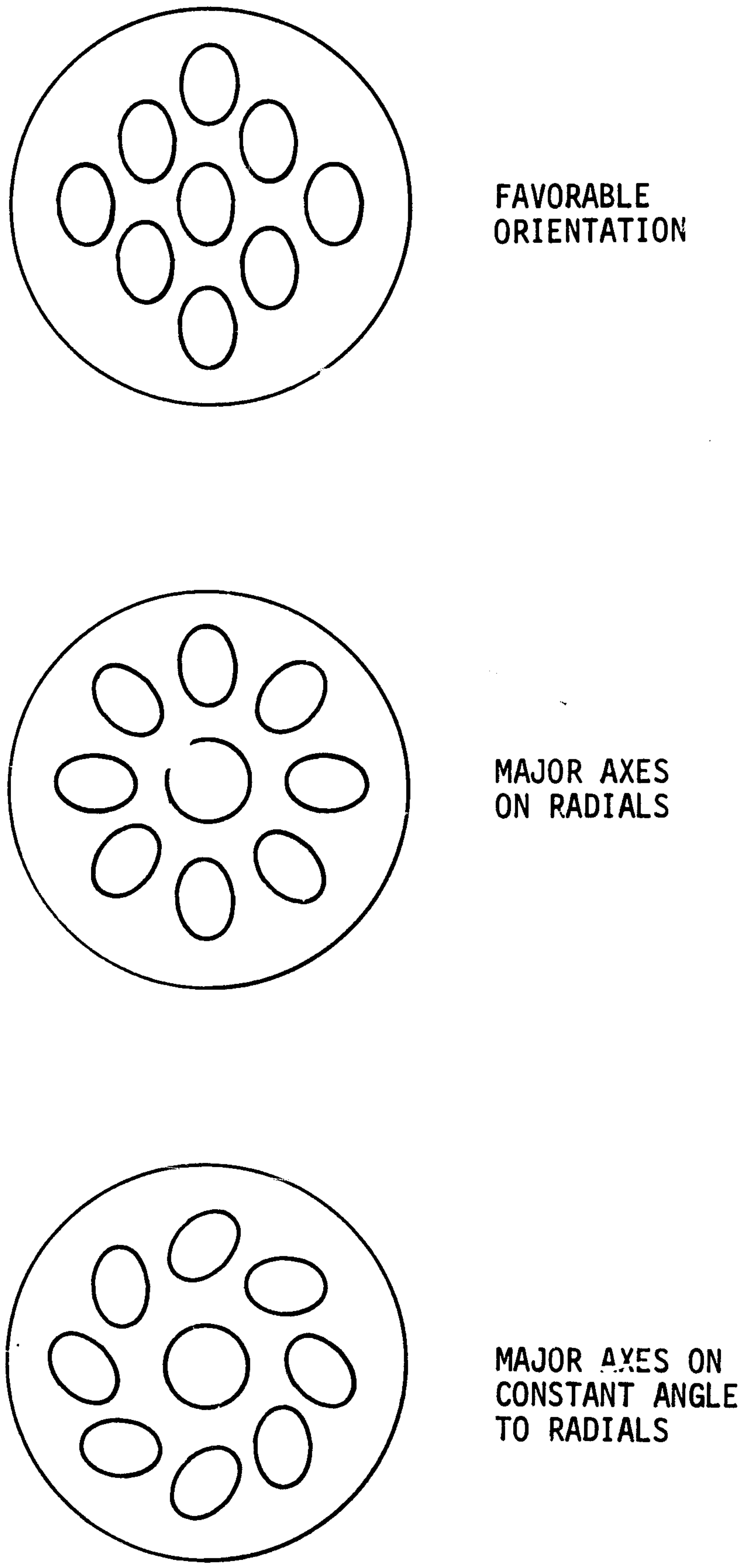


Figure 4-15 POLARIZATION ELLIPSE TOPOLOGY ON FAR-FIELD SPHERICAL CAPS

Figure 4-16a and 4-16b[39] show measured pattern/axial-ratio responses of two conical spiral antennas with one rotated 90° about its beam axis. It may be noted that axial ratios at ±70° off axis are ≤4 dB for both elements. (These are 2 arm spirals designed for somewhat less than cardioid beamwidths.)

Figure 4-17 shows the broad beamwidth pattern transverse to the line of the 4-element array of the conical spirals of Figure 4-16a & 4-16b. The exceedingly low axial ratios out to ±80° are evident. In particular, the axial ratios at ±70° are ≤2 dB as a result of the crossed ellipse element beams. Figure 4-18 shows the narrow beam cut of the 4-element array. Axial ratios over the entire main beam are <0.5 dB and under 2 dB over all sidelobes out to ±55°. The array dimensions for this antenna are shown in Figure 4-10. The crossed ellipse technique can be applied to broader cardioid beam con-spiral elements to extend the extra-low axial ratio performance to ±90° in the wide beam plane. Thus, a desired low axial ratio sectoral beam can be realized, although with some complexity of design.

The requirement for good circular polarization at large angles from beam center is an obvious requirement for electronically steered arrays. For linear arrays, it is also relevant, since the hemispherical coverage requirement can be met by using a broad fan beam, steered in one axis only. For such an application, circular polarization is required at all angles where the antenna has usable gain. It is also, of course, important in the switched-beam and electronically scanned beam array configurations. These categories are discussed in Sections 5 and 6.

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(39) Unpublished Measurements by Philco-Ford Corporation, Newport Beach, California. (Private Communication with Mr. W.G. Scott.)

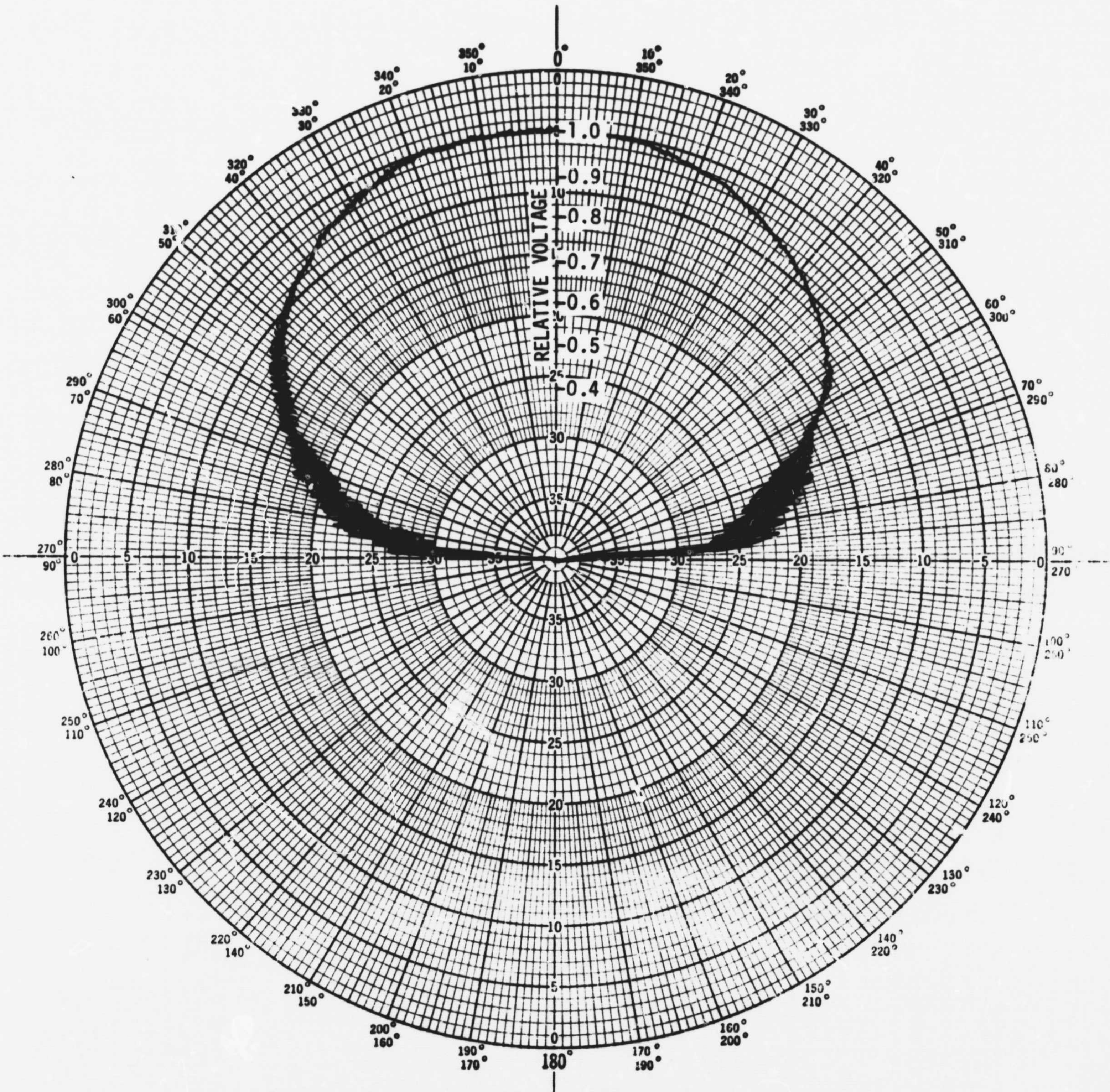

Figure 4-16a CONICAL SPIRAL ANTENNA PATTERN; C = 0°

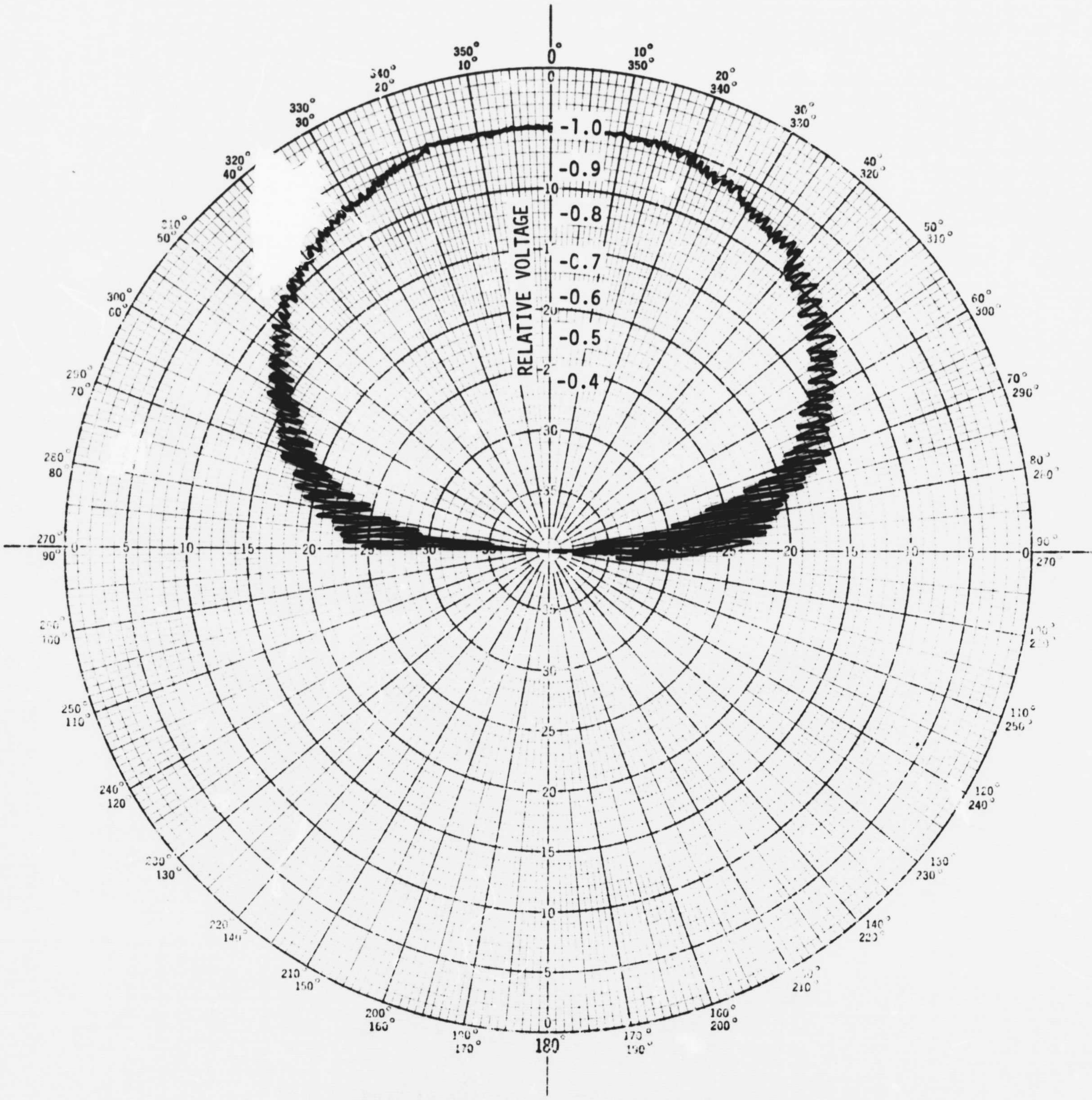

Figure 4-16b CONICAL SPIRAL ANTENNA PATTERN; C = 90°

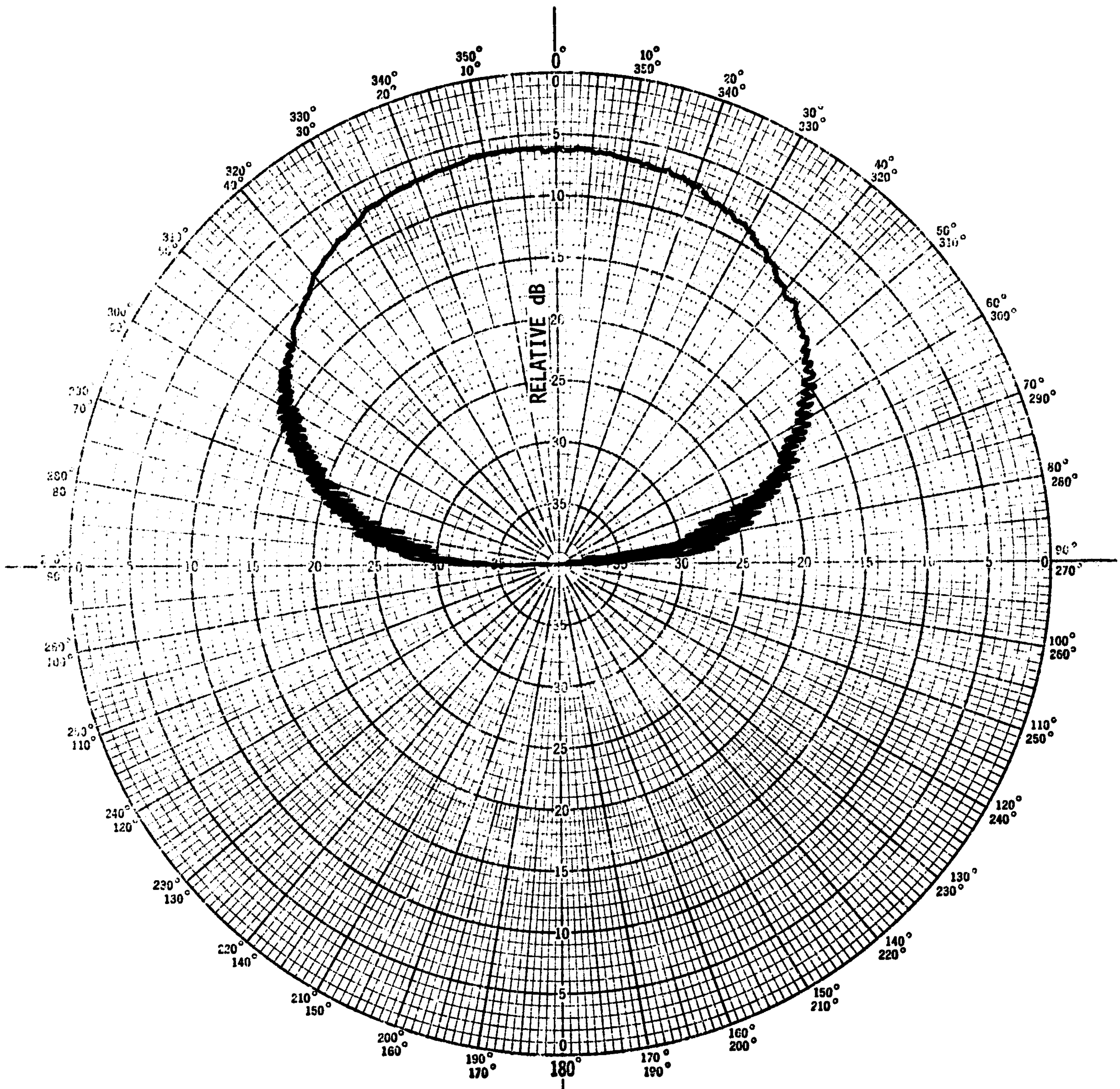

Figure 4-17 AZIMUTH PATTERN OF 4-ELEMENT ARRAY, F = 1000 MHz, S = 1.0λ

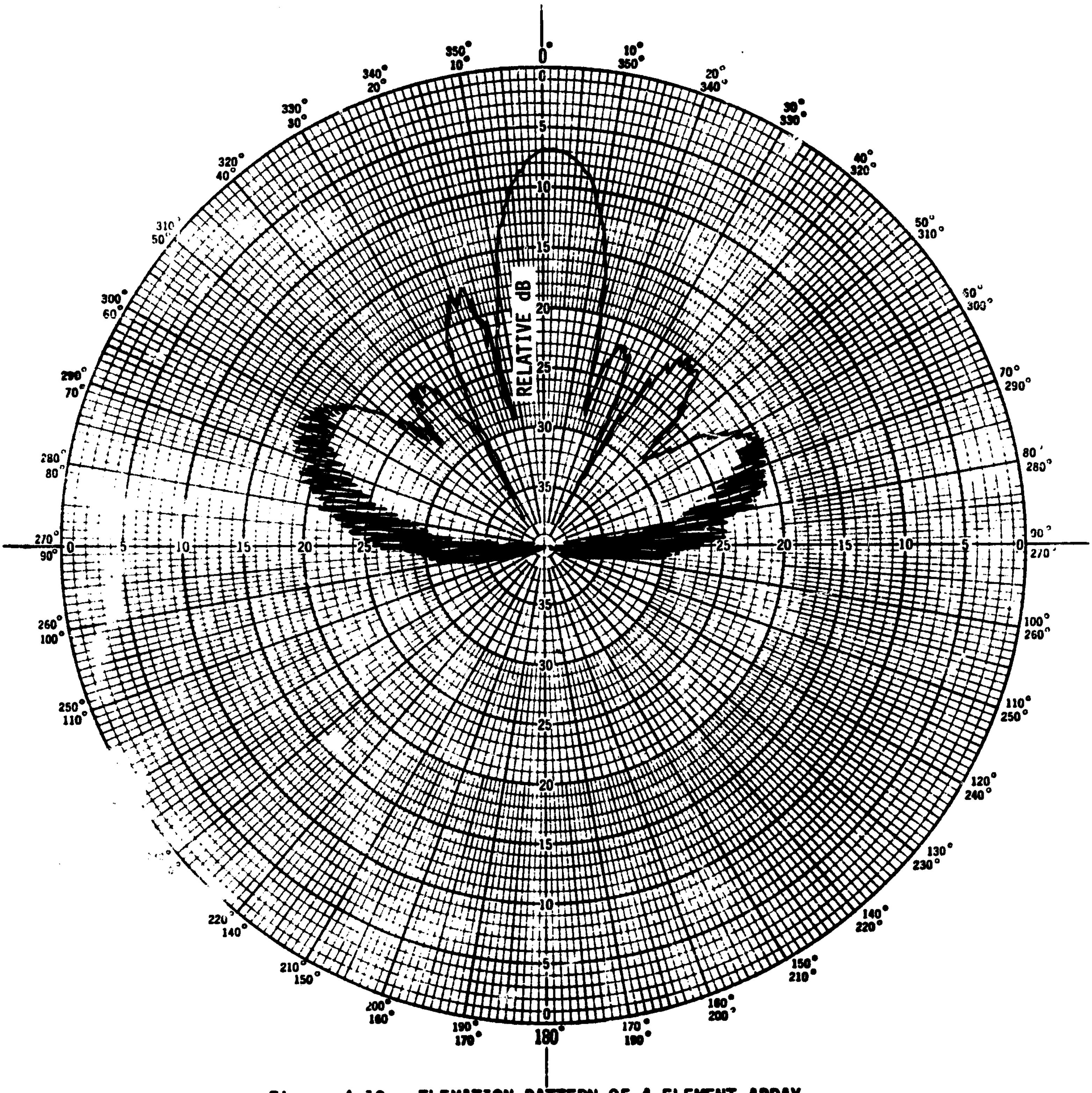

Figure 4-18 ELEVATION PATTERN OF 4-ELEMENT ARRAY, F = 1000 MHz, S = 1.0λ

For weather protection, a conformal-type radome could be used over the array for shipboard installation as shown in Figure 4-10. While the multiple elements in an array inherently offer the potential of developing a difference pattern for automatic tracking, the fan beam configuration discussed is primarily associated with single-axis slaving to shipboard inertial references. Thus, while the use of fan beams implies certain limitations in gain (Section 2.5), it is possible to generate the desired beam shape.

4.6 CROSSED YAGI-UDA ARRAY

4.6.1 General

The crossed Yagi-Uda array is an end-fire parasitic antenna which consists of a single driven antenna and closely coupled parasitic elements as shown in Figure 4-19.

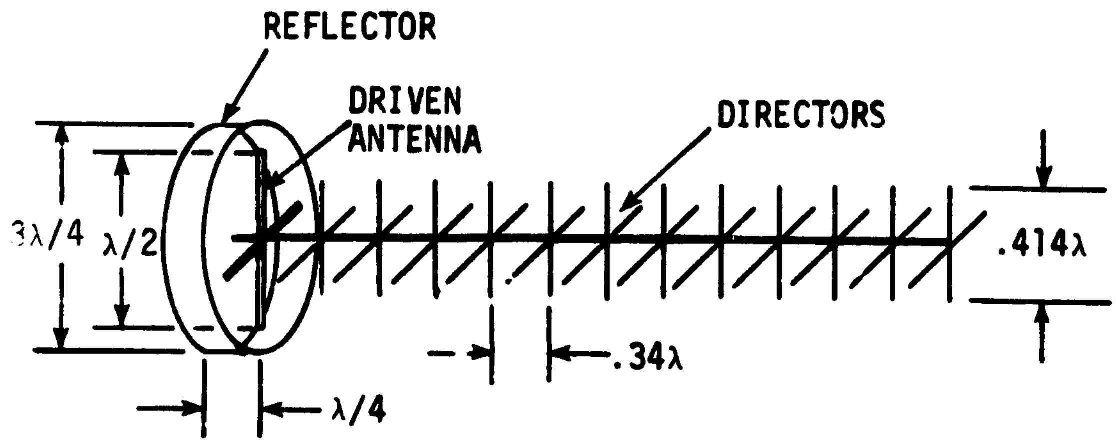

Figure 4-19 CROSSED YAGI-UDA ARRAY

The antenna is circularly polarized by feeding the crossed-dipole driven antenna elements in phase quadrature. The function of the directors is to increase the antenna gain. Each director receives its power parasitically from the preceding director and in turn radiates a portion of its power. The incremental contribution from the directors is progressively less along the array. The practical limit to the total number of elements is 12 or less although arrays of thirty or forty elements have been built. The director length and spacing as presented in Figure 4-19, represents a good compromise on gain, input impedance, front-to-back ratio, and bandwidth for the antenna.

4.6.2 Performance

Representive performance reported is presented in Figure 4-20 where the antenna gains and beamwidths are plotted against the number of directors.

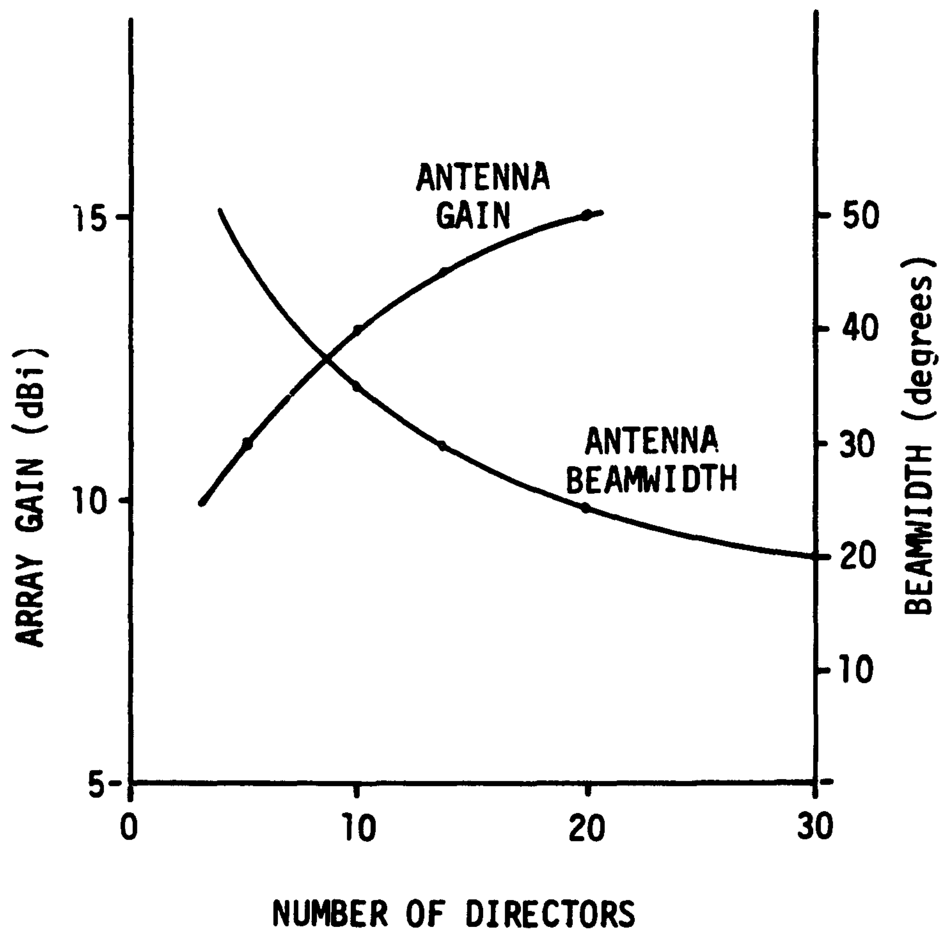

Figure 4-20 CROSSED YAGI-UDA ARRAY GAIN AND BEAMWIDTH VS NUMBER OF DIRECTORS

With the crossed dipole elements excited by a turnstile feed circuit, the antenna can be made to perform monopulse tracking by the utilization of the sum and different pattern through the respective circuit ports. However, more than 2 crossed sets of elements may be required to excite the difference beam mode so required. This is not difficult to do, however.

For shipboard use, the Yagi-Uda antenna must be kept short, because it gets flimsy as its length increases, and this would limit the useful gain range to approximately 10 dB.

4.7 HELICAL RADIATORS

4.7.1 General

A gain of 3 to 15 dB can be easily achieved by a helical design of proper dimensions. The electrical performance characteristic of the helical antenna is very predictable. The axial mode radiation is obtained when the helix circumference is $\sim$ 1 wavelength. The helix dimensions are noncritical in nature; and, as a result, the axial mode radiation is easily generated in practice. For this reason, the helical antenna is considered to be one of the simplest typesof antenna to make. The helical antenna also possesses excellent broadband properties, including desirable pattern, impedance and polarization characteristics over a relatively wide frequency range up to 1.67:1 of the mid frequency.

4.7.2 Performance Characteristics

The more important performance characteristics such as pattern, beamwidth, directivity and terminal resistance may be accurately computed from the respective formulas presented below:

Pattern $$E = \left(\sin \frac{\pi}{4n}\right) \frac{\sin\left(\frac{n\psi}{2}\right)}{\sin\left(\frac{\psi}{2}\right)} \cos \phi \qquad (4\text{-}9)$$

Beamwidth (half-power) $$B = \frac{52}{C_\lambda \sqrt{n S_\lambda}} \text{ deg} \qquad (4\text{-}10)$$

Beamwidth (first nulls) $$B = \frac{115}{C_\lambda \sqrt{n S_\lambda}} \text{ deg} \qquad (4\text{-}11)$$

Gain $$G = 10\, C_\lambda^2\, n\, S_\lambda \qquad (4\text{-}12)$$

Terminal Resistance $\quad R = 140\ C_\lambda$ ohms $\quad$ (4-13)

Axial Ratio $\quad AR = \frac{2\ n + 1}{2\ n}$ $\quad$ (4-14)

where

$$\psi = 360° \left[S_\lambda\ (1 - \cos\phi) + \frac{1}{2n} \right]$$

n = number of turns of helix

ϕ = angle off axis in radians

C_λ = circumference in free-space wavelengths

S_λ = spacing between turns in free-space wavelengths

The geometry of a typical helical design is shown in Figure 4-21.

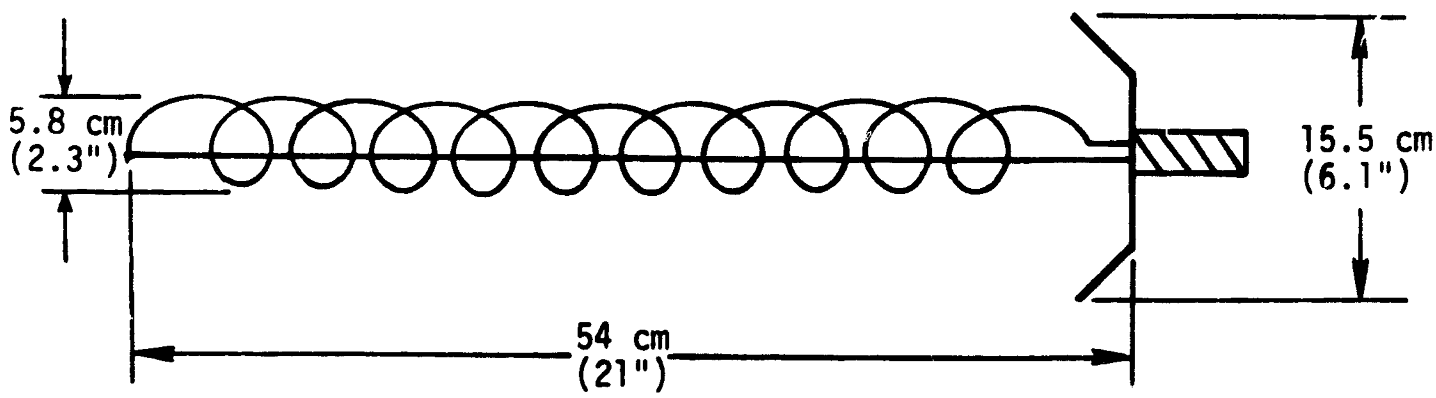

Figure 4-21 TYPICAL HELICAL ANTENNA GEOMETRY

The antenna operates over a 15.5 cm diameter cup ground plane. The diameter of the helix is 5.9 cm. The helix length is 53 c.. The pitch angle is 14°.

In terms of the antenna dimensions, the calculated antenna performance is presented below:

Beamwidth (half-power)	= 32.7°
Beamwidth (first nulls)	= 72.4°
Gain	= 14.5 dB
Terminal Resistance	= 140 ohms
Axial ratio	= 0.34 dB

Helical antennas are popular for the reception of spacecraft telemetry signals. They can be made cheaply and simply and can be easily protected from the shipboard environment by enclosing the helix in a fiberglass cylinder.

Helices are also very commonly used in 3- or 4-element planar arrays for spacecraft tracking.

4.8 LOG-PERIODIC CROSSED DIPOLE ANTENNAS

4.8.1 General

The log-periodic crossed dipole is a type of circularly-polarized, broad-band antenna which consists of an array of crossed dipoles with lengths and spacing arranged in a log-periodic manner. The antenna may be excited by feeders from back to front, as shown in Figure 4-22a). In this configuration, the antenna becomes its own balun. A 90° hybrid provides the necessary phase quadrature currents for the dipoles. Radiation of the antenna is endfire in the direction of the shorter element. The bandwidth of the antenna is determined by the size of the elements used, where the highest and lowest frequencies are determined respectively by the shortest and longest dipole. The practical limit of gain for the log-periodic, crossed-dipole antenna is approximately 9 dBi. The gain is relatively constant over the design bandwidth.

4.8.2 Design Considerations

A comprehensive analysis of the antenna and step-by-step design procedure are presented by Carrel.[40] The number of elements is determined by the design parameter, τ; the number of elements is increased as τ increases. The antenna size is determined by boom length (the distance between the smallest and largest element) which depends on σ. The boom length increases as σ increases. The parameter relationships are presented in Figure 4-22b). In L-band design, the antenna would be much larger than a helix or Yagi-Uda antenna. Complexity and cost would be higher. The side- and backlobe characteristics are less desirable.

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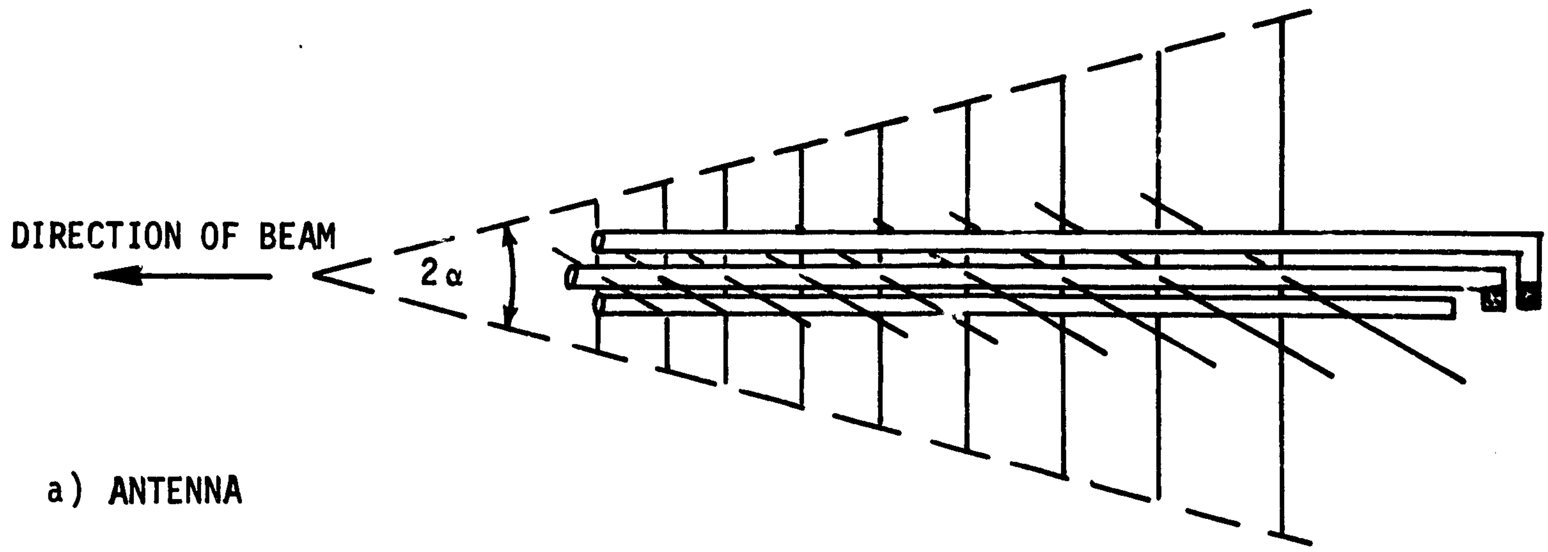
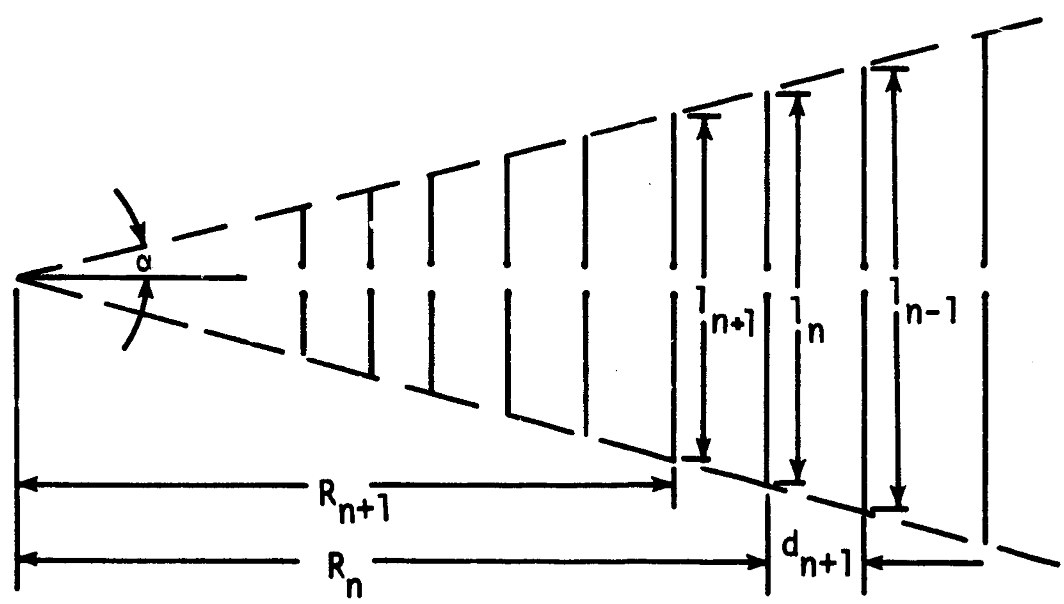
[40] Carrel, R. "The Design of Log-Periodic Dipole Antennas." IRE International Convention Record 9, 1961.



a) ANTENNA



b) DESIGN PARAMETER RELATIONSHIPS

$$\frac{R_n}{R_{n+1}} = \frac{l_n}{l_{n+1}} = \tau$$

$$\frac{d_n}{2l_n} = \sigma$$

$$h_n = \frac{l_n}{2}$$

Figure 4-22 LOG-PERIODIC CROSSED DIPOLE ANTENNA

The log-periodic antenna is primarily used in those applications where its broad bandwidth can be used to advantage. In the application under consideration here, the disadvantages noted above mean that it is not a serious contender.

4.9 CONICAL AND RECTANGULAR HORNS

4.9.1 General

A horn antenna is a flared end on a hollow conductor transmission line or waveguide. The flare acts as an impedance transformer to convert from the waveguide impedance to that of the radiating wave in free space. The throat of the horn also acts as a mode filter, allowing only a single mode to propagate from guide to aperture. Therefore, if the flare angle is not too large, all modes other than the dominant will be eliminated at the aperture. Rectangular, circular and elliptical waveguides and flared horns have been used, but rectangular waveguide is most common and will be assumed in this section.

4.9.2 E-Plane Horns

Figure 4-23 illustrates E-plane sectoral horns. Figure 4-24 from Silver[21] shows the relationship of E-plane aperture width (in wavelengths) versus E-plane pattern beamwidths. An E-plane 10 dB beamwidth of 160° is indicated with a one-half wavelength E-plane aperture width.

An optimum wide angle sectoral pattern with low side lobes requires a large aperture width, contrary to the apparent trend in Figure 4-24. The wide patterns of small aperture horns never have steep "edges" to the patterns, and considerable radiation spills beyond the intended beam sector. As illustrated in Figure 4-25, from reference 21, the use of an in-phase circular arc aperture of increasing electrical width produces an increasingly better approximation to an idealized pre-shaped sectoral pattern with less and less radiation beyond the sector angle

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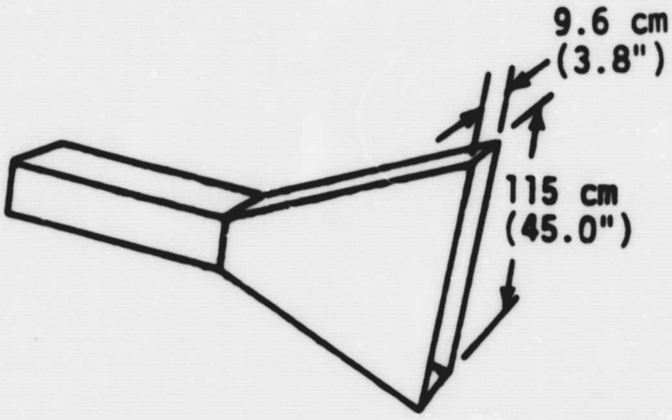
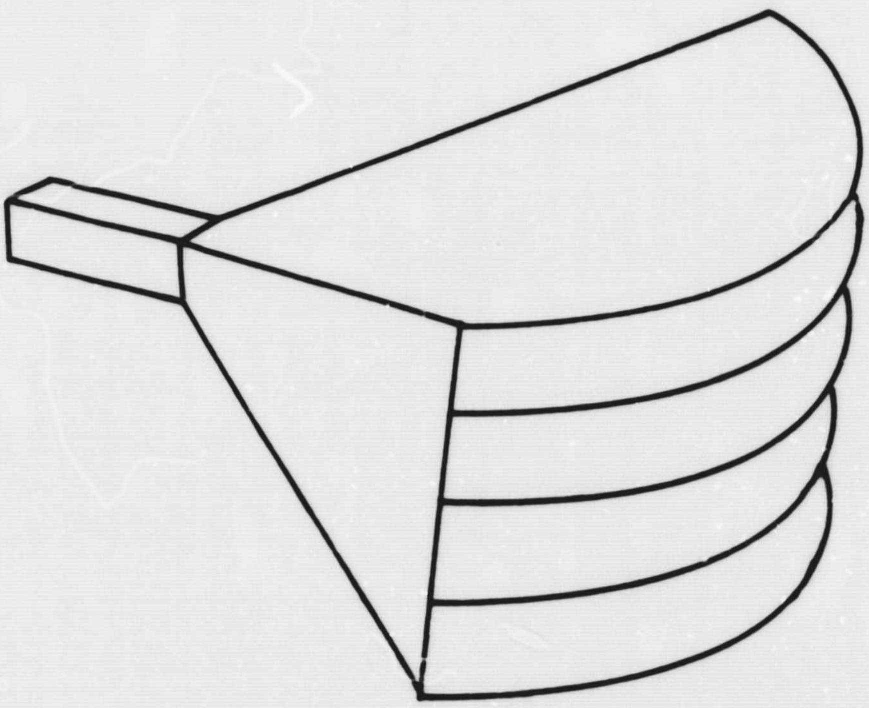
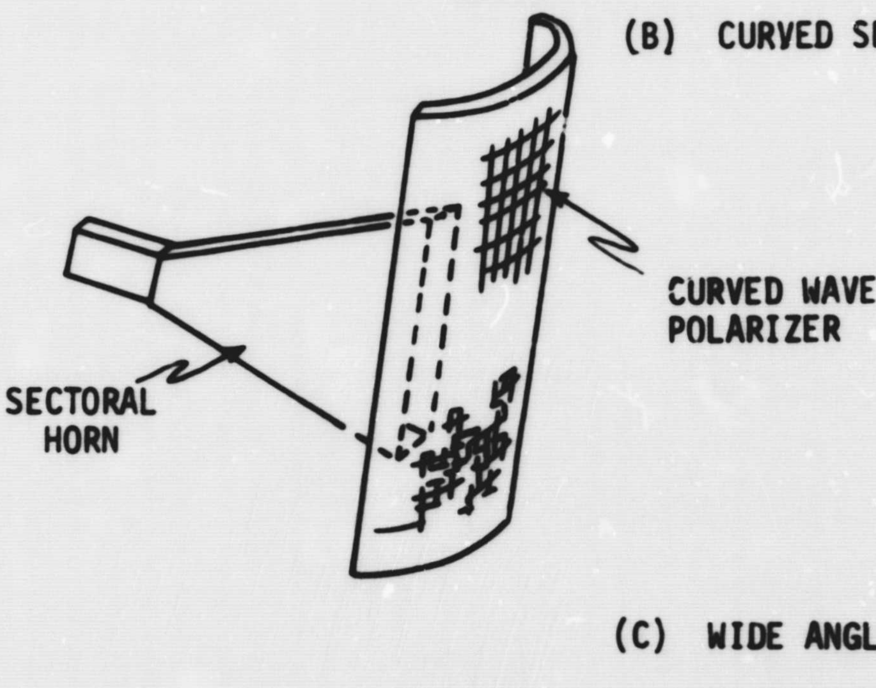
(21) Silver, S. Op. Cit.



(A) SECTORAL HORN



(B) CURVED SECTORAL HORN



(C) WIDE ANGLE CIRCULAR POLARIZED HORN

Figure 4-23 SECTORAL HORNS

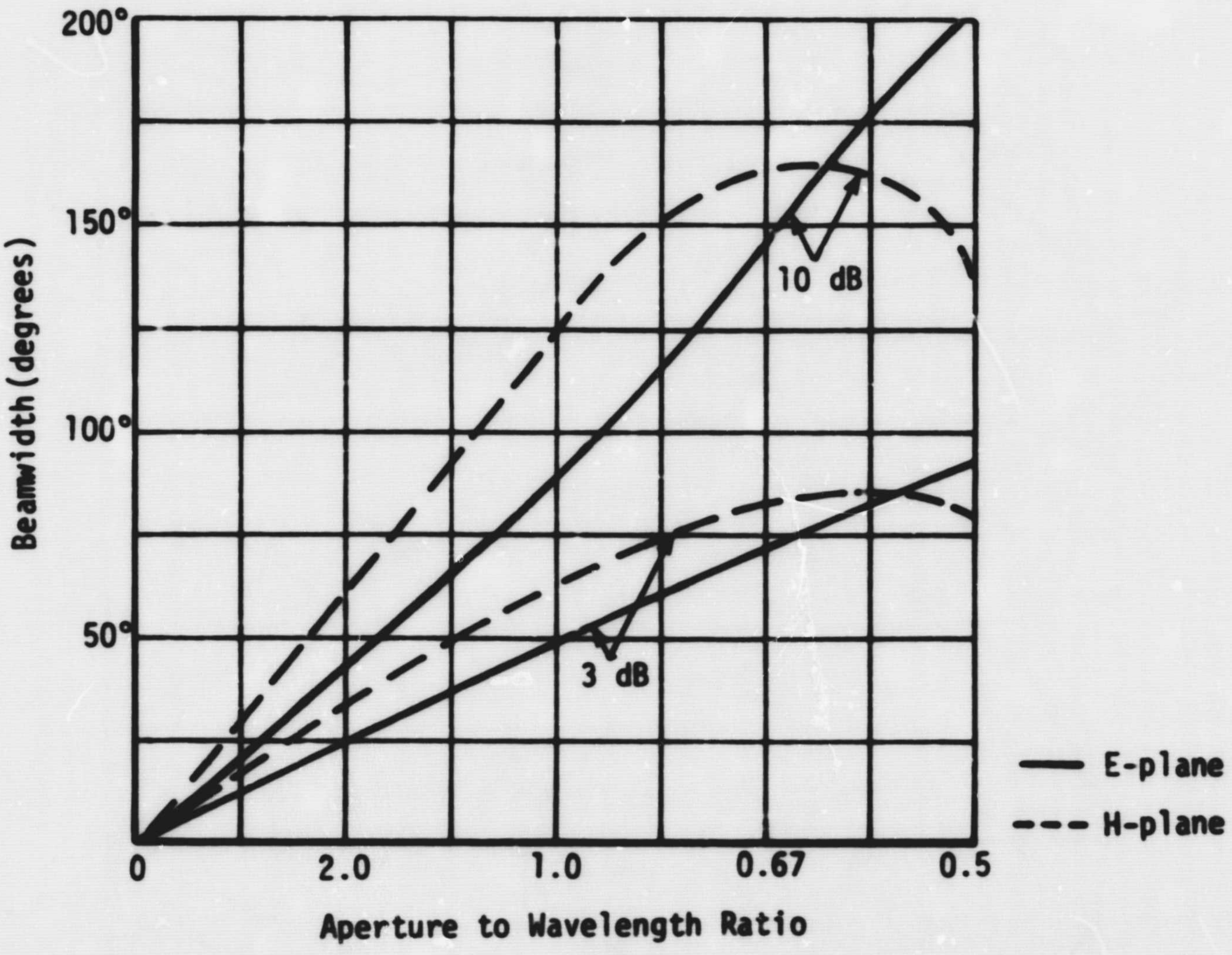


Figure 4-24 RELATION BETWEEN THE APERTURE DIMENSION AND THE 3 dB AND 10 dB WIDTHS OF THE RADIATION PATTERN OF RECTANGULAR WAVEGUIDE

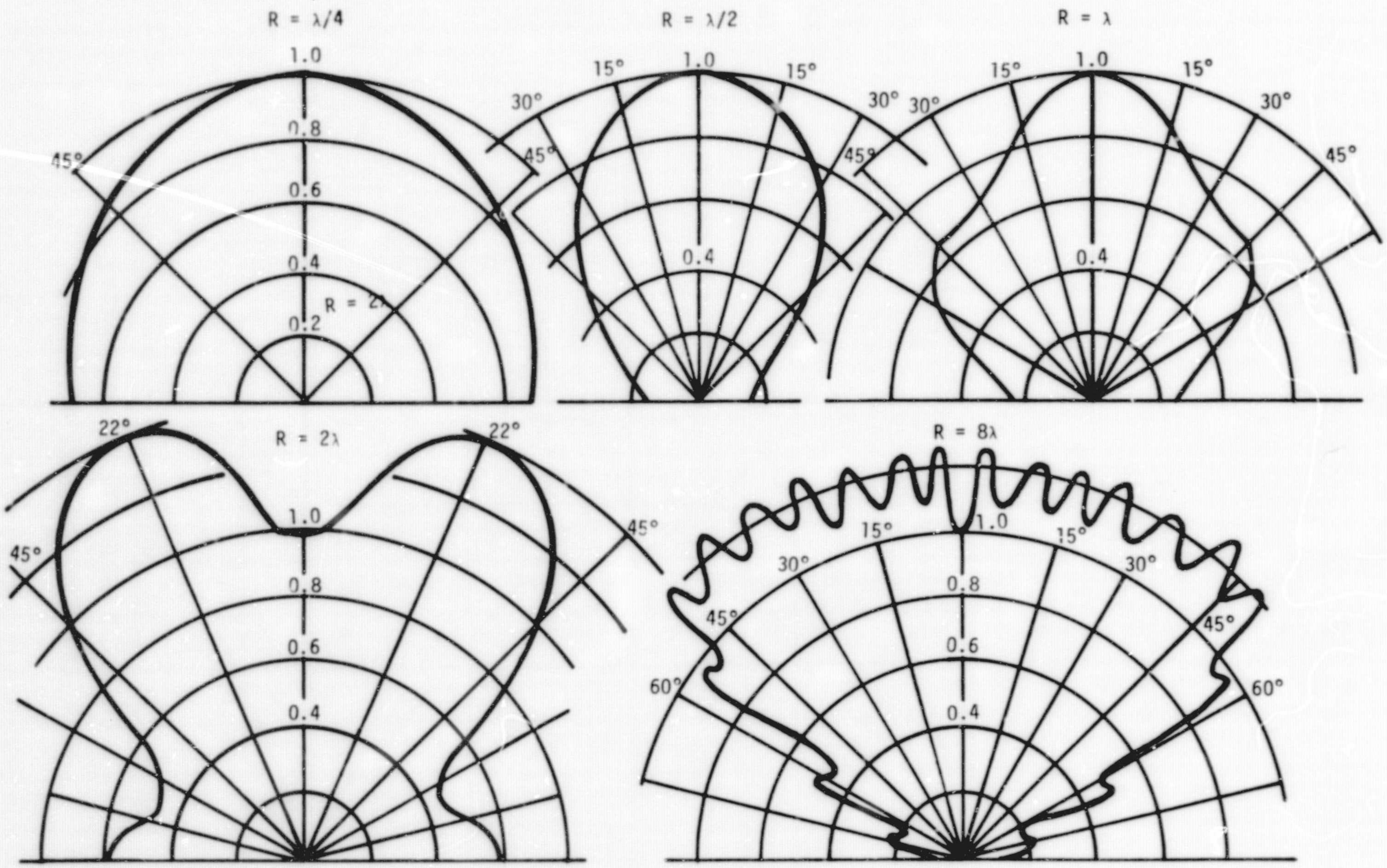


Figure 4-25 SECTORAL PATTERNS OF 120° ARC APERTURE HORN

subtended by the aperture arc. Thus a tradeoff exists--one may use quite small apertures for wide beamwidths (with spillover) or quite large circular apertures for wide beamwidths (with very little beam spillover and fairly uniform pattern amplitude over the beamwidth). This tradeoff is illustrated in comparing the conventional sectoral horn of Figures 4-23a) with 4-23b), where the circular arc aperture is realized in a compound horn. Both horns have the same sectoral beamwidths but not the same beam efficiencies (spillover).

4.9.3 Circular Polarization

Circular polarization can be achieved for horn antennas by several means, but extremely low axial ratios are difficult to maintain at wide pattern angles such as ±>60° off axis.

Figure 4-23c) illustrates the use of a wave polarization converter screen spaced a short distance from the sectoral horn aperture to achieve circular polarization. This converter screen changes the linearly polarized wave radiated from the horn aperture to one of circular polarization.

Lerner[41] shows a version of such a screen composed of three photo-etched sheets and a dielectric honeycomb core. The conductive foil etched on the plastic sheets forms a grid of thin strips (shunt inductance) interlaced with rows of isolated rectangles (shunt capacitance). With the incident wave E-field oriented at forty-five degrees to the grid pattern, circular polarization can be obtained. By use of three such sheets, a wide-band impedance match can be obtained. Lerner

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(41) Lerner, D.S. "A Wave Polarization Converter for Circular Polarization." IEEE Trans AP, January 1965.

reported that a curved cylindrical plate, mounted over a sectoral horn, produced an insertion loss of less than 0.5 dB with an axial ratio under 1.7 dB. A further advantage of the polarizer is that it can serve as the horn radome.

Horns are very commonly used as feeds for high-gain antennas, where their easily controllable characteristics and relatively uniform aperture power distribution make possible the design of highly efficient systems. Obtaining circular polarization from a horn antenna is relatively complex, and is a considerable disadvantage, as the simple horn is otherwise small, lightweight and inexpensive to produce.

4.10 THE SHORT BACKFIRE ANTENNA

4.10.1 General Description

One very attractive single-beam antenna candidate for gains on the order of 12 to 14 dB is a relatively new design referred to as a "short backfire" antenna. It has the very desirable characteristics of high efficiency and simplicity of construction. The short backfire antenna shown in Figure 4-26 consists of two planar reflectors of different diameters, separated by one-half wavelength, forming a shallow leaky cavity resonator with the radiation beam normal to the small reflector. The antenna, originated by Ehrenspeck,[42] is fed by a dipole at the midpoint between the two reflectors and yields useful gains of 10-15 dBi. Typical values for the reflector diameters at 1540 MHz are 2.0 wavelengths (39.0 cm) and 0.5 wavelengths (9.8 cm), as shown in Figure 4-27, for 14 dBi gain and 35° beamwidth. The addition of a quarter-wave rim on the larger reflector reduces back radiation from -20 dB to below -30 dB. The pattern and gain bandwidth have been measured to be 17% in one example, although the natural impedance match bandwidth is only 3-5% for VSWR's under 2.0 and might have to be broadened with a matching network to provide a low VSWR at the transmit frequency of 1640 MHz (i.e., for 6.4% bandwidth).

The major interest in this antenna is due to the fact that its aperture efficiency is approximately 70%, a figure very difficult to obtain in parabolic reflectors of this small diameter (2.0 wavelengths). With the high aperture efficiency goes exceptionally low sidelobe levels, helping to reduce multipath and ship structure scattering losses. Figures 4-28 and 4-29 give some measured[43] beamwidth, sidelobe and gain parametric

(42) Ehrenspeck. "The Short Backfire Antenna." Proc. IEEE (Correspondence). Vol. 53, pp. 1138-1140, August 1965.

(43) Dod, L.R. "Experimental Measurements of the Short Backfire Antenna." Goddard Space Flight Center, Greenbelt, Md., Report S-525-66-480, October 1966.

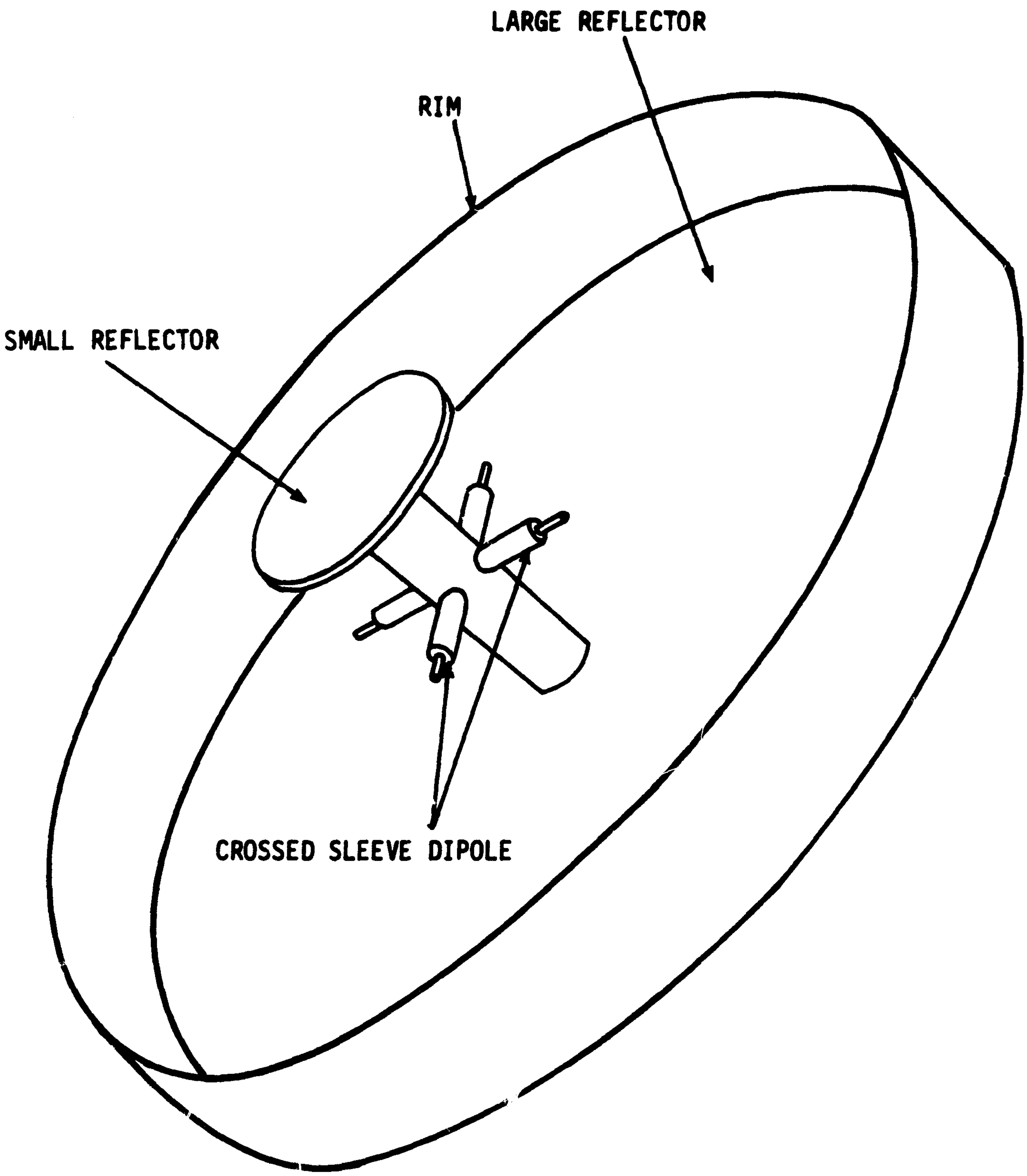


Figure 4-26 SHORT BACKFIRE ANTENNA

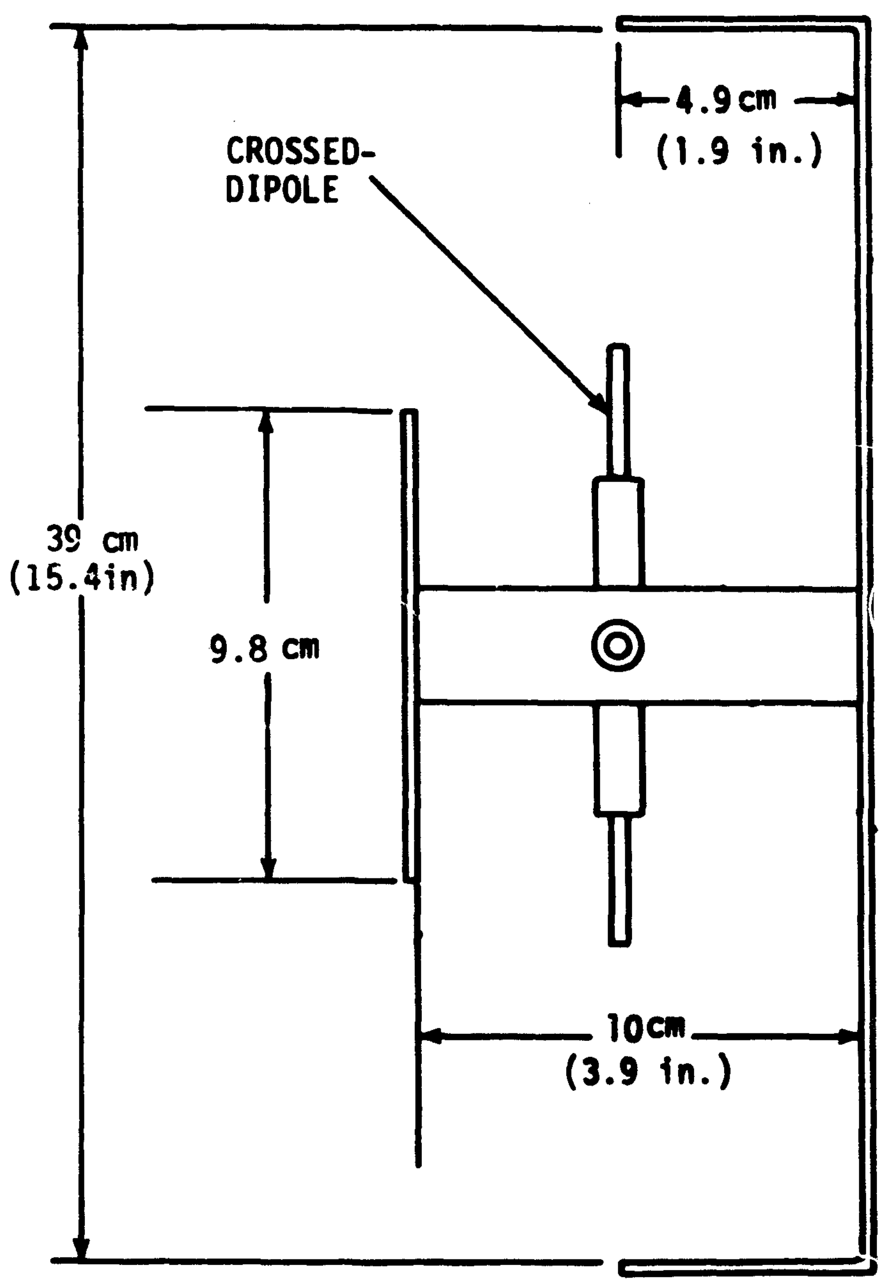
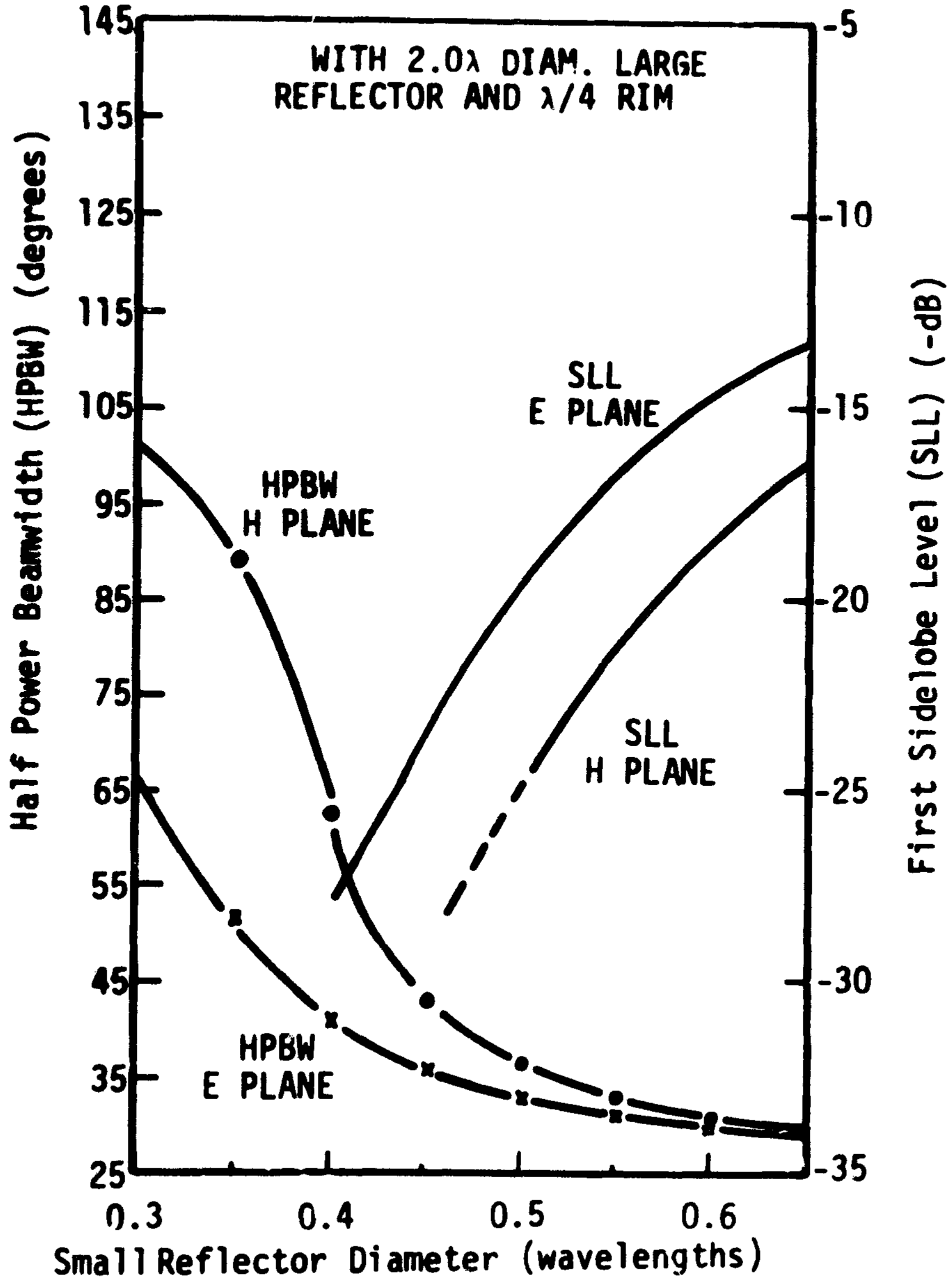
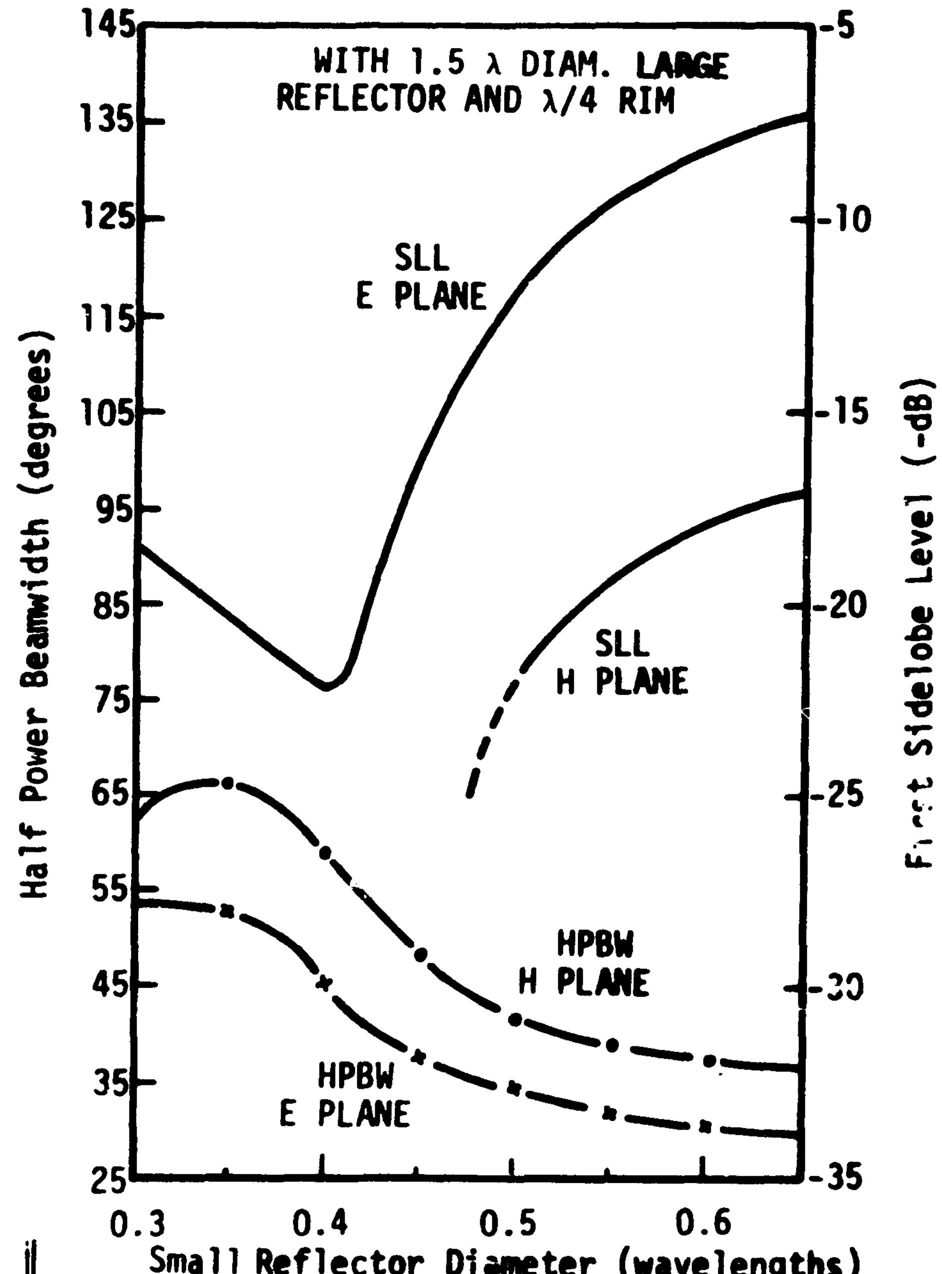


Figure 4-27 DIMENSIONS OF SHORT BACKFIRE ANTENNA



(A) Half Power Beamwidth and Sidelobe Level Versus Reflector Diameter 2.0λ Diameter Reflector with λ/4 Rim.



(B) Half Power Beamwidth and Sidelobe Level Versus Reflector Diameter 1.5λ Diameter Reflector with λ/4 Rim.

Figure 4-28 SHORT BACKFIRE ANTENNA BEAMWIDTH DATA

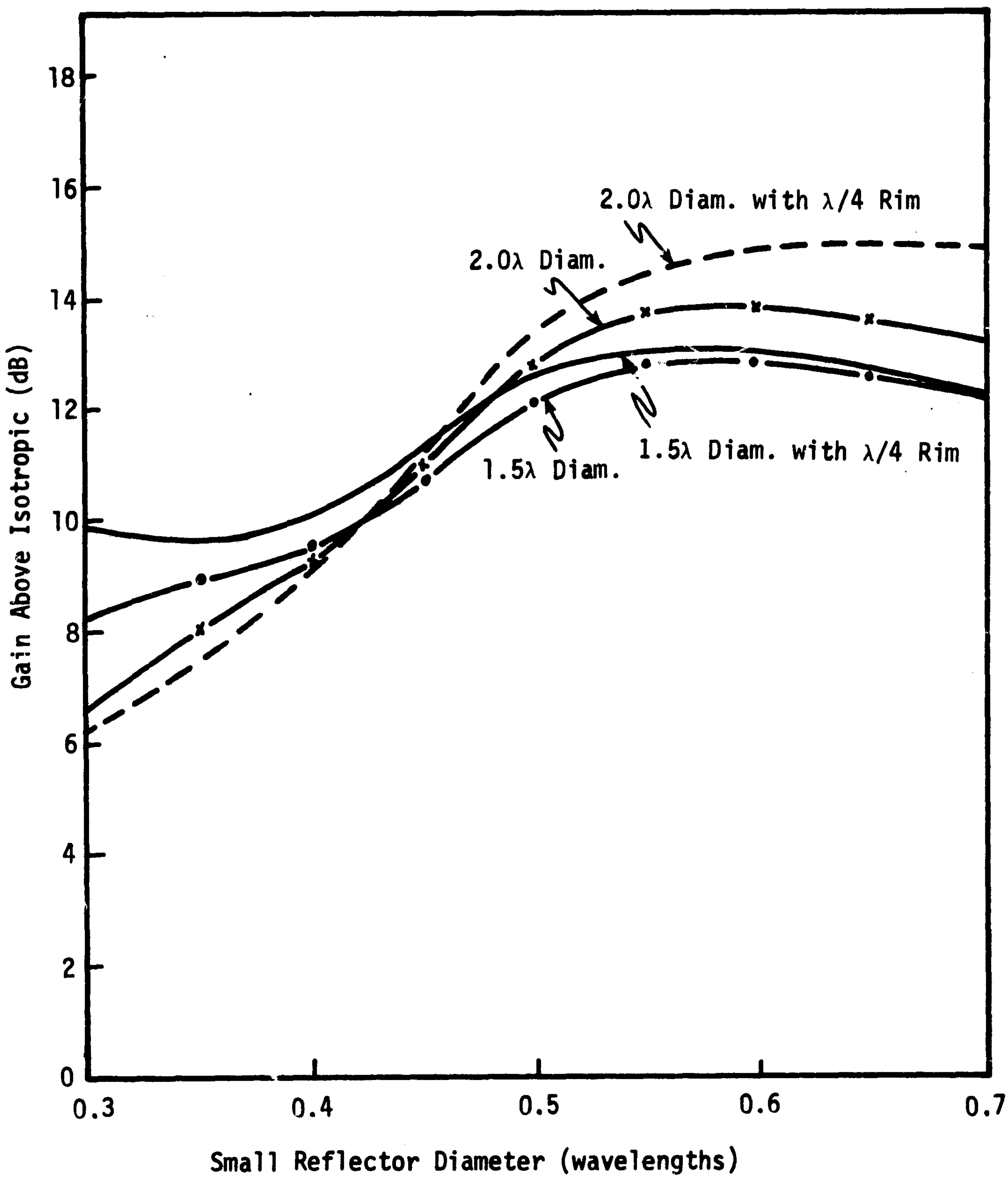


Figure 4-29 SHORT BACKFIRE ANTENNA GAIN DATA

data. In addition, it is quite simple in construction and not sensitive to dimensional tolerances.

4.10.2 Principles of Operation of the Short Backfire Antenna

The name "short-backfire antenna" is misleading since the operating mode for the backfire antenna is characterized by multiple reflection of electromagnetic waves between the two plane reflectors, with a standing wave field distribution along the antenna axis. This condition requires that the reflectors be separated by n $\lambda_0/2$, where n is an integer, such that the tangential E fields vanish at their surfaces. The stored energy in the near-field zone is concentrated in the region between the reflectors along the antenna axis, and is radiated through the aperture plane. A short-backfire antenna with reflector separation of 0.5 wavelength is the shortest possible dimension for antenna configurations in this category.

The aperture distribution has been shown to be that of a cavity resonator; it is thus cosinusoidal in both E and H planes. The dipole acts merely as an exciter for the cavity field and may be replaced by another structure of similar dimensions, such as a loop, for example.

4.10.3 Tracking Capability of the Short Backfire Antenna

In order to obtain monopulse sum (Σ) and difference (Δ) beam performance, the circular symmetry of the reflectors permits the use of a multiple crossed dipole feed. A pair of orthogonal dipoles can be excited in

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(44) Chen, R.M., Nyquist, D.P., Lin, J.L. "Radiation Fields of the Short-Backfire Antenna." IEEE Trans. on Antennas and Propagation. pp. 596-597, September, 1968.

quadrature as a turnstile feed (0°, 90°, 180°, 270° phasing) to produce circular polarization in the axial beam. Although a 4-arm log-spiral antenna can also be excited with a 720° phase progression (0°, 180°, 360°. 540°) so as to produce a circularly polarized conical, point nulled Δ beam simultaneously with a circularly polarized Σ beam, this is not in general true of 4-arm structures of other shapes, such as the 4 poles of the crossed dipole. In fact, a remarkable proof was recently published[45] for n-arm antennas exhibiting "n-fold symmetry", i.e., antennas having n identical arms of any shape, regularly disposed about an oriented axis, a, such that a rotation through an angle $2\pi/n$ about the axis, a, leaves the structure invariant, each arm being carried into congruence onto the next one. The n arms are fed equally, but with an increasing phase increment of $2\pi/n$ for the Σ (axial) beam pattern and $4\pi/n$ for the Δ (point nulled) pattern. The theorem states that for such n-arm antennas, in the vicinity of the antenna axis a, circular polarization is necessarily obtained in the Σ pattern, regardless of the shape of the antenna arms if $n > 3$, and in the Δ pattern if $n > 4$. Thus, a 5, 6, 7, 8,-... arm linear crossed-dipole feed will automatically produce circular polarization in both the Σ and Δ patterns. However, in the case of a 4-arm antenna, circular polarization is not obtained automatically in the Δ channel excitation but, if achieved at all, is obtained only because of the nature of the antenna structure; i.e., as for the 4-arm log-spiral antenna.

Consequently, for the short-backfire antenna, a set of 3 crossed dipoles ($n = 6$) will suffice to produce circularly polarized Σ and Δ monopulse modes as in the 4, 6, 8, ... arm spiral antennas.[46] A relatively simple

(45) Crout, P.D., "The Determ[illegible]on of Antenna Patterns of n-Arm Antennas by Means of Bicomp[illegible] Functions." IEEE Trans. on Antennas and Propagation, pp. 686-6[illegible], September, 1970.

(46) Deschamps, G.A. and Dys[illegible], J.D. "The Logarithmic Spiral in a Single Aperture Multi-mode Antenna System." IEEE Trans. Antennas and Propagation, pp. 90-96, January, 1971.

stripline hybrid matrix, a form of the Butler Matrix[47] can be used to simultaneously excite the Σ and Δ modes for monopulse operation, which may then be converted to a single channel monopulse receiver configuration, as discussed in Section 3.5.3.

For shipboard operation, the antenna itself can be readily enclosed by a honeycomb fiberglass radome for weather protection, and still be a very small package (e.g., a cylinder 10 cm long and 40 cm in diameter).

The Short Backfire Antenna would thus appear to be a most suitable candidate for a shipboard antenna. Its extremely high efficiency gives it relatively high gain, but not at the expense of a narrow beamwidth (which would complicate the tracking function). It is capable of providing an axial null pattern for tracking without excessive complication. The small size, ease of construction and compact, rugged shape are also advantages for shipboard use.

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(47) Butler, J. and Lowe, R. "Beam Forming Matrix Simplifies Design of Electronically Scanned Antennas." Electronic Design, p. 170, April, 1961.

4.11 TURNSTILE ON GROUND PLANE

4.11.1 General

The turnstile antenna is a design where two dipoles are arranged orthogonal to one another and are excited with phase quadrature current of equal magnitude. The radiation of the antenna is circularly polarized in the axial direction from the crossed elements.

A turnstile antenna may be used to provide a unidirectional pattern by operating it over a ground plane as shown in Figure 4-30.

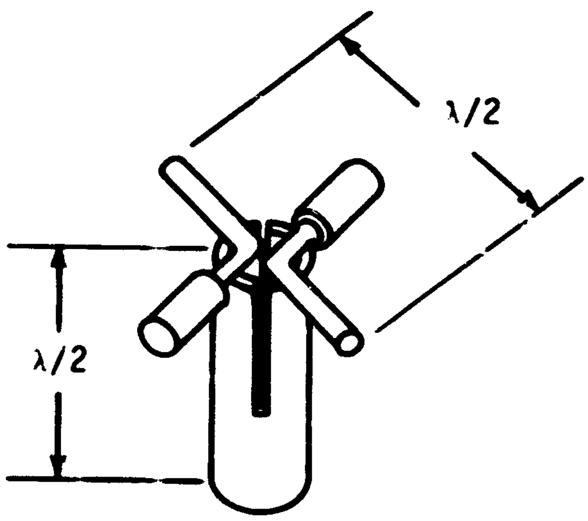

Figure 4-30 TURNSTILE ANTENNA

In this design, the dipole elements are excited by a split balun and the necessary phase quadrature current is obtained by making one dipole slightly shorter than the other. This antenna is very simple to design and relatively inexpensive to make. The antenna pattern shape is approximately [illegible], where Θ is the angle off the antenna axis. The peak gain of the antenna is slightly greater than 5 dB. The impedance bandwidth is slightly less than 10 percent. The antenna design is well suited to the subject, shipboard applications as a very low gain radiator, and is very simple to manufacture. The antenna may also be used as an array element.

4.11.2 Circular Polarization Characteristics

With the turnstile on ground plane radiator shown in Figure 4-3, the axial ratio of the circular polarization degrades with increasing angle Θ away from the beam axis. This is shown in Figure 4-31.[48] For example, at $\Theta = 52$ degrees, the axial ratio is 5.1 dB. From Figure 4-32[49] it is seen that, for a satellite axial ratio of 3.0 dB, a polarization decoupling loss of 0.9 dB results.

A way of reducing the wide angle axial ratio of the turnstile is shown in Figure 4-33,[48] where the turnstile arms are curved above the ground plane. It should be noted that the axial ratio at $\Theta = 52°$ is 2.6 dB, a considerable improvement over the conventional flat turnstile. Satellite polarization decoupling loss would then be less than 0.4 dB.

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(48) Scott, W.G. "Special Antenna for a Moon Capsule." Electronics. November 16, 1962. pp. 46-49.

(49) "The Microwave Engineers Handbook and Buyers Guide." Horizon House, Dedham, Mass. February 1966.

(48) Scott. Op. Cit.

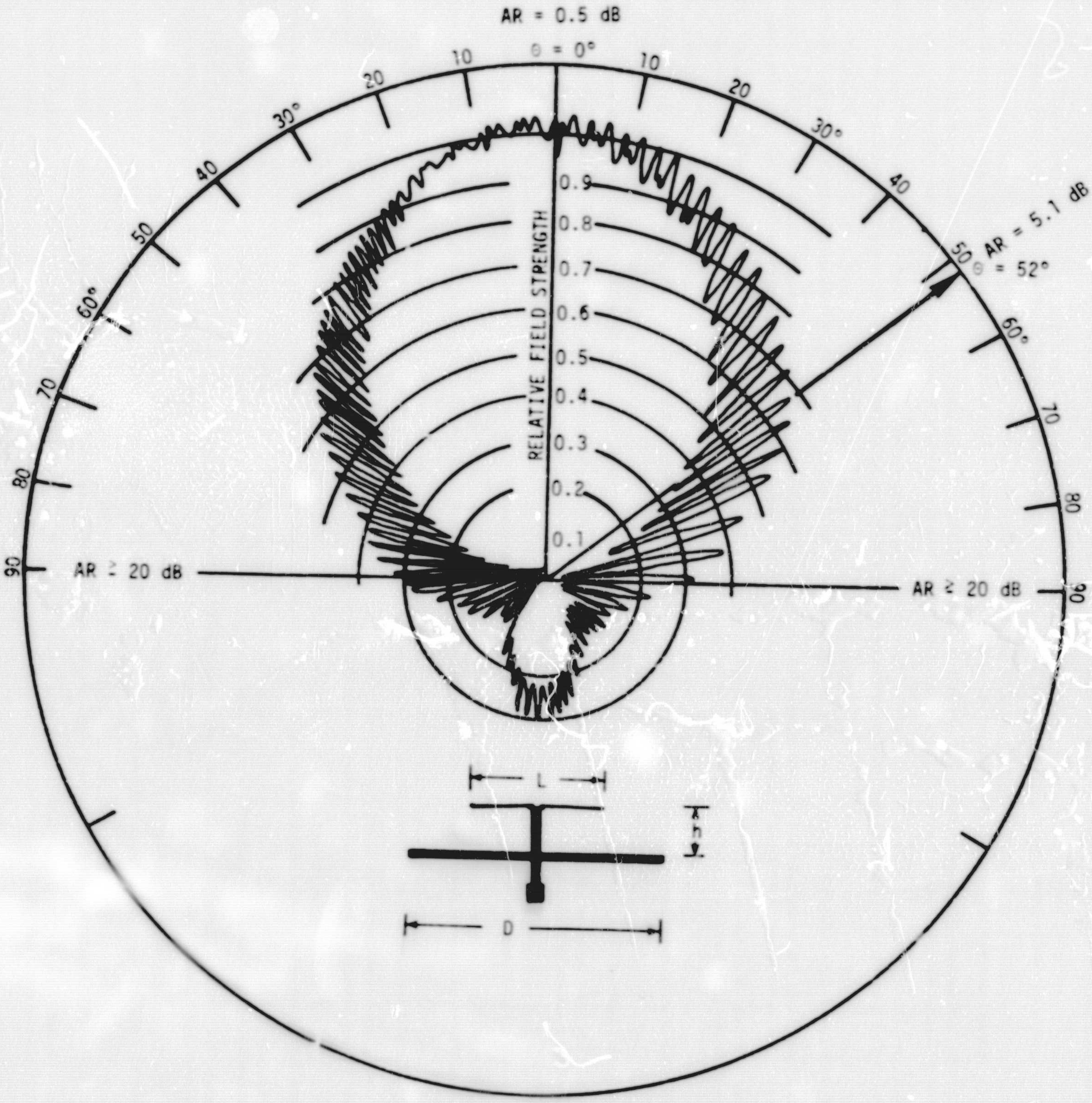
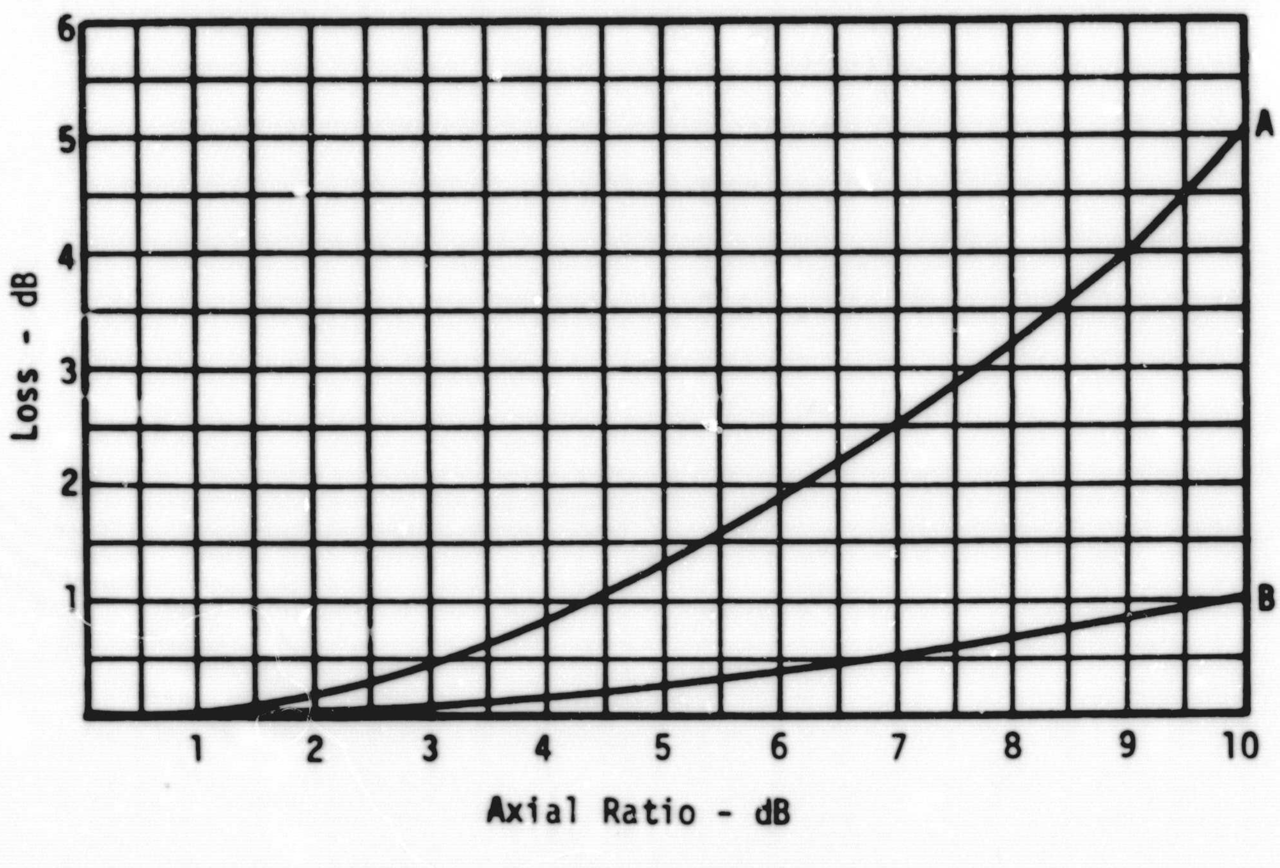


Figure 4-31 TURNSTILE OVER-GROUND PLANE AXIAL RATIO OFF-AXIS



A. Same Sense--same axial ratios

B. Same Sense--one circularly polarized

Figure 4-32 MAXIMUM POLARIZATION LOSS BETWEEN TWO ELLIPTICALLY POLARIZED ANTENNAS

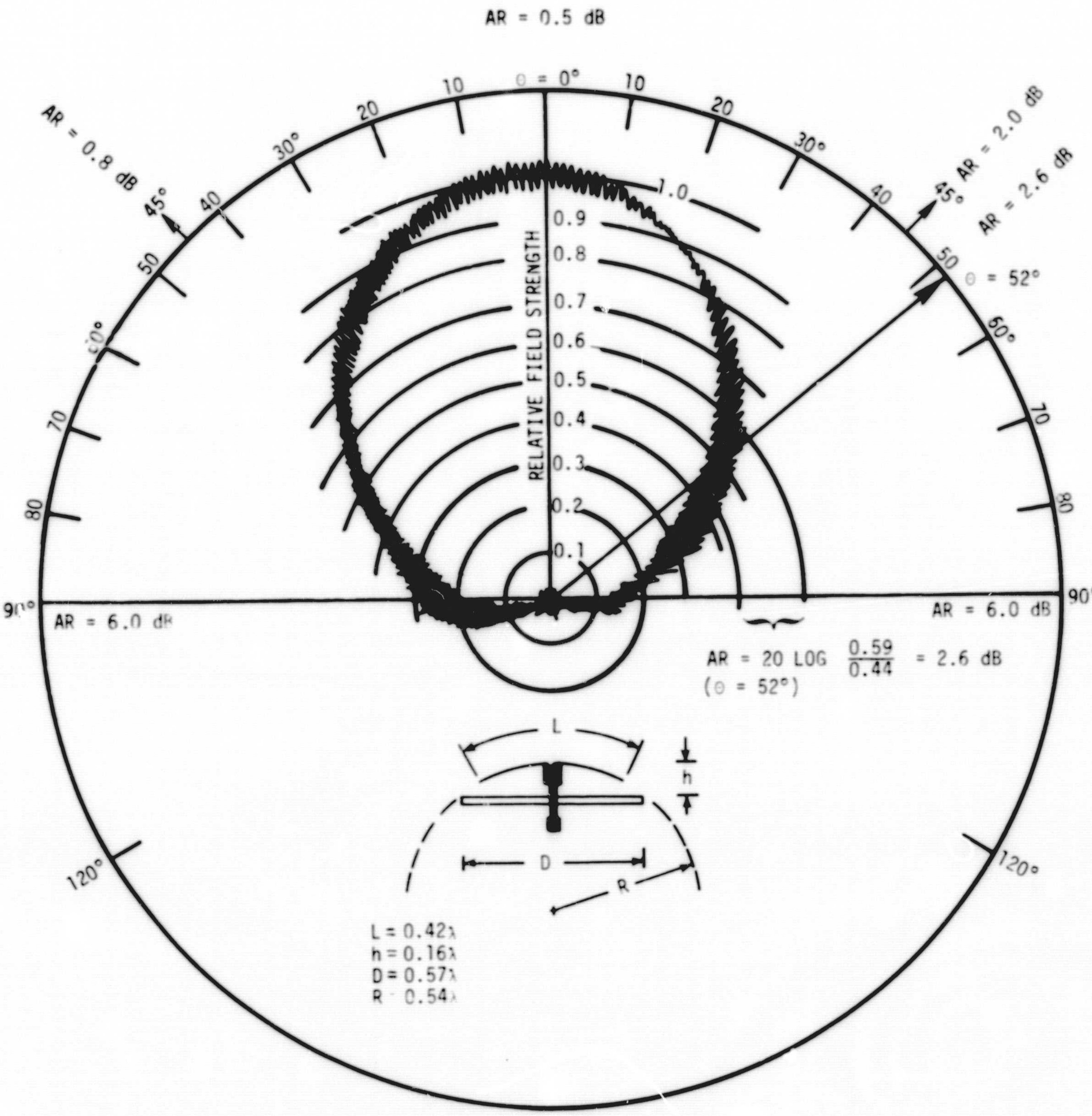


Figure 4-33 CURVED TURNSTILE FOR WIDE ANGLE CIRCULAR POLARIZATION

4.12 CONICAL LOG-SPIRAL

4.12.1 General Characteristics

The conical log-spiral is a circularly polarized, unidirectional radiation antenna with extremely broad bandwidth (i.e., frequency independent). However, the wide angle circular polarization pattern performance of this antenna rather than its broad band properties, is of interest in the maritime satcom application. The antenna is constructed by wrapping balanced equiangular spiral arms on a conical surface fed at the tip of the cone. The sense of rotation of the circularly polarized field is determined by the direction of winding of the spiral arms. Several variations of the design are available. The primary variation is in terms of the number of spiral arms. As an example, the two arm configuration provides a single lobe of radiation along the axis of the cone in the direction of its apex, whereas the four arm configuration may be excited to radiate two modes of radiation, an axial beam and a point-nulled conical beam. The four arm configuration is thus capable of generating a pattern null which can be used to provide for dual channel monopulse tracking. The antenna peak gain, however, is generally less than 6 dBi for the axial beam and less than 5 dBi for the conical beam. For hemispherical or greater coverage, the requirement is easily met with a two arm design.

4.12.2 Design Parameters

The performance of the conical log spiral antenna can be appropriately specified in terms of the five parameters given below:

1) Included cone angle, 2θ
2) Armwidth angle, δ
3) Spiral rate, α
4) Base diameter, D
5) Apex diameter, d

These parameters are shown in Figure 4-34. The originator of the conical spiral antenna, Dyson,[50] has found that by increasing δ to 45°, half-power beamwidths up to 180° can be obtained. However, the polarization axial ratio at the beam edges degrades with increasing beamwidth, so that for a 180° beamwidth, axial ratios of 6.0 dB typically exist.

4.12.3 Improved Circular Polarization for Conical Log-Spirals

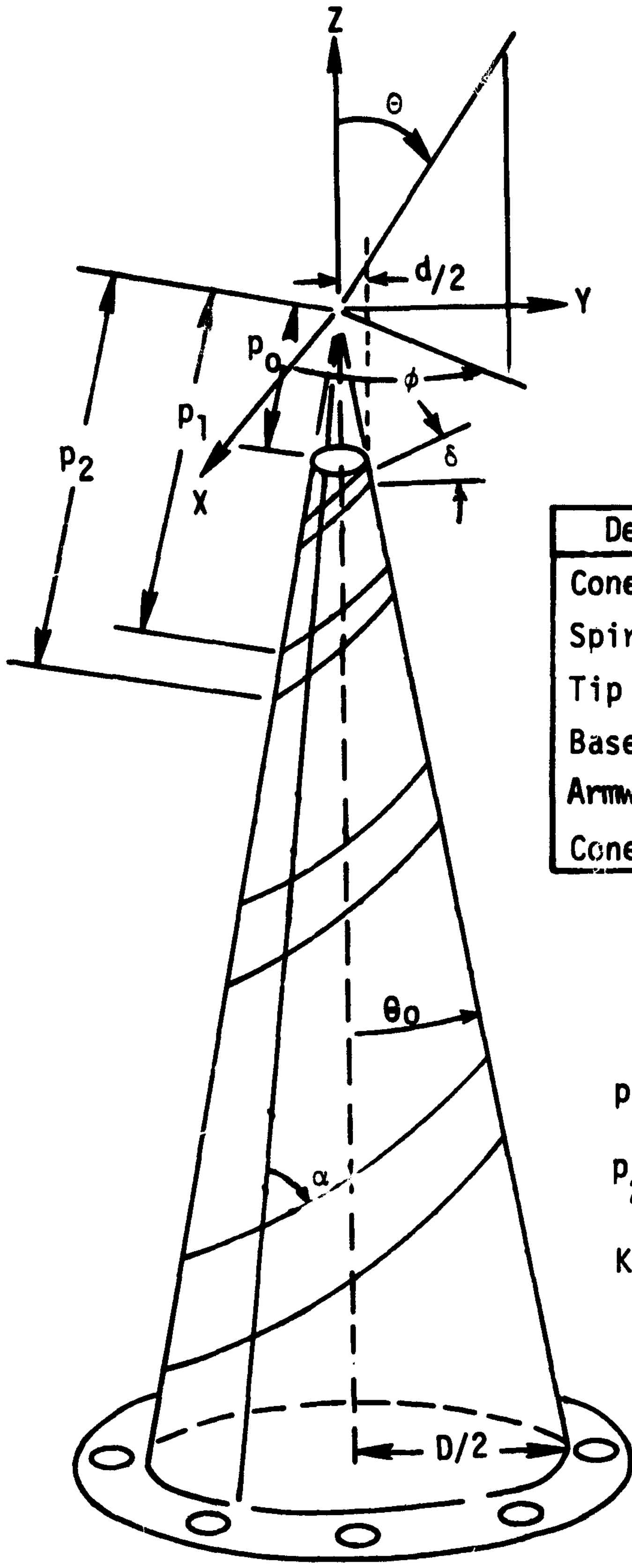
For beamwidths of 180°, Scott, et al[51] have shown that 4 and 8 arm conical spiral antennas provide improved axial ratios. Figure 4-35 and 4-36 show patterns of 4 and 8 arm spirals with hemispherical beams. (The oscillation on the basic pattern is a measure of the axial ratio along the pattern.) For the 4 arm spiral, the desired approximation to a hemispherical beam with one sense of circular polarization is within ±0.85 dB of constant amplitude and under 2.7 dB axial ratio over the hemisphere. An 8 arm design provides even better axial ratio performance. Here the axial ratio stays under 1.7 dB and pattern amplitude is constant to within ±1.3 dB over the hemisphere.

4.12.4 Conical Log-Spiral Configuration Aspects

The multi-arm (n-arm) designs are driven with equal amplitude per arm and $2\pi/n$ phase progression per arm for the axial beam. A hybrid strip-line network is required to feed these n terminals. Such networks can

(50) Dyson, J. "New Circularly Polarized Frequency Independent Antennas With Conical Beam or Omnidirectional Patterns." IRE Trans. on Antennas and Propagation. July 1961.

(51) Scott, W. G., Ermatinger, C.E., Westerman, C.W. and Harrington, V.L. "Wide-angle Circularly Polarized Antenna Techniques for Spacecraft." IRE 9th Conference on Aerospace and Navigational Electronics. Oct. 1962.



Design Values at 1540 MHz	
Cone Half Angle	$\Theta_0 = 7.5°$
Spiral Angle	$\alpha = 45°$
Tip Diameter	$d = 1.0$ cm
Base Diameter	$D = 8.6$ cm
Armwidth Angle	$\delta = 90°$
Cone Length	= 23 cm

$$p_1 = p_0\, e^{(\sin\Theta_0/\tan\alpha)\,\phi}$$

$$p_2 = K p_1$$

$$K = e^{(-\sin\Theta_0/\tan\alpha)\,\delta}$$

Figure 4-34 HEMISPHERICAL COVERAGE CONICAL SPIRAL ANTENNA

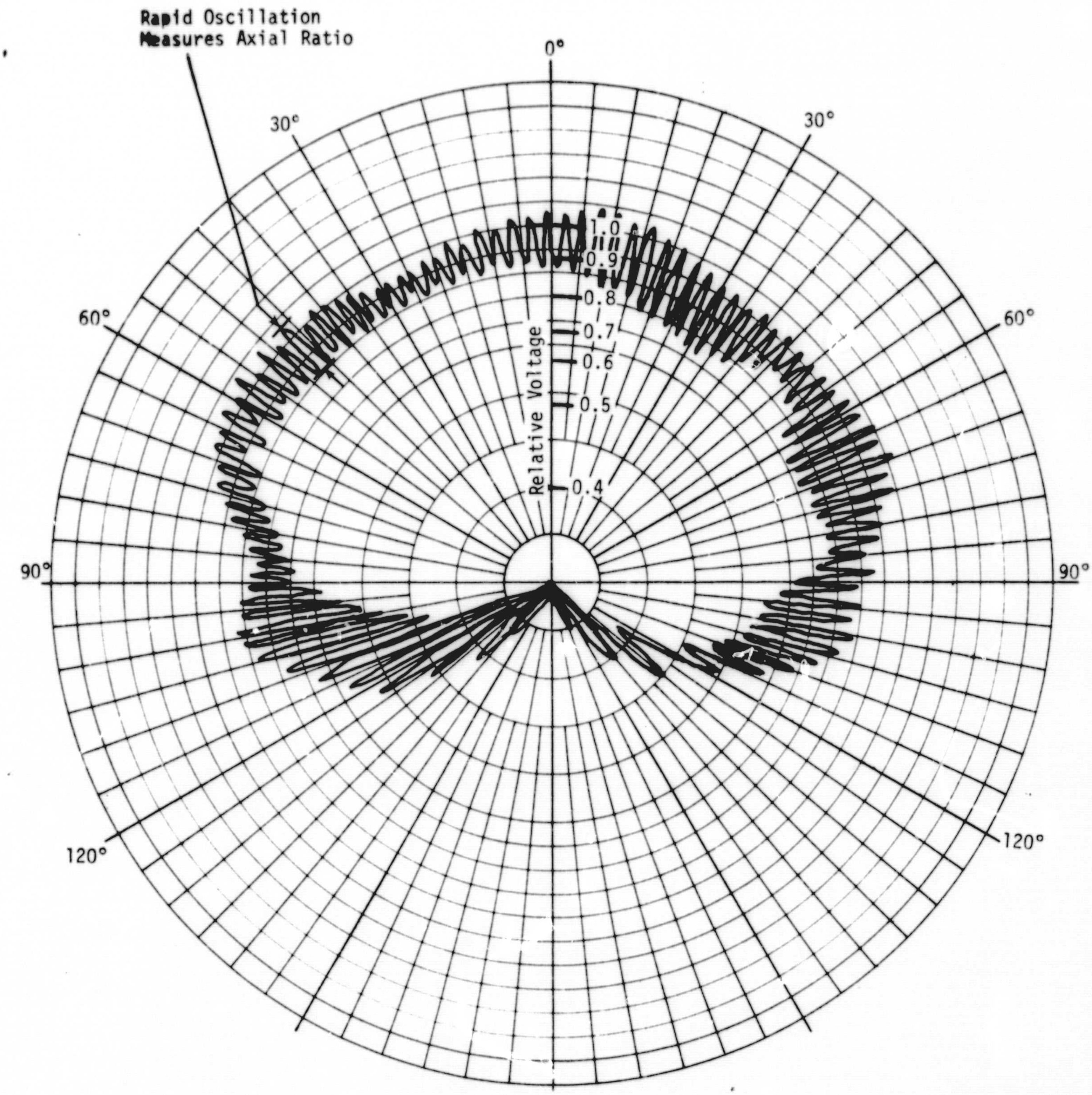


Figure 4-35 4-ARM CONICAL SPIRAL PATTERN

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be small and rugged.

An alternate hollow conical beam shape can also be obtained from a given n-arm conical spiral if $h \geq 4$. This beam is obtained by an equal amplitude excitation with phase increments of $4\pi/n$ per arm. The beamwidth of the axial beam and the tilt angle of the center of the hollow conical beam are controlled primarily by the spiral arm pitch angle α. Increasing α (tighter winding) gives narrower axial beams and tighter (narrower) hollow conical beams. Increasing the number of arms produces smoother patterns with lower wide angle axial ratios, but increasing n requires a feed network of increasing complexity. Four, six and eight arm antennas are, however, practical and have been used. Another benefit of increasing n is an extremely low level of back radiation, i.e., less than-20 dB. This can be advantageous for minimizing unwanted scattering from shipboard structures.

In general, the connection between n coaxial feed cables and the n spiral arms at the cone tip is the only delicate part. A small conformal plastic radome over the spiral with interior foam filler would be adequate for environmental protection. Spacing of the radome from the spiral can be designed to minimize deleterious effects of ice coatings on antenna impedance and VSWR. In its wide-angle form, the conical spiral antenna is eminently suitable for use as an array element in linear (fan beam) arrays and in electronically steered or switched arrays.

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4.13 CAVITY-BACKED SPIRALS

4.13.1 General

A cavity-backed spiral is a type of broad-band antenna which utilizes the effect of a cavity to provide a unidirectional beam perpendicular to the antenna aperture. The antenna is circularly polarized according to the winding sense of the two or more printed circuit spirals. Nearly all the spirals in use are either Archimedean or equiangular spirals. The antennas are built with equal spiral and cavity diameters. A reasonable spiral diameter for the subject application is 15 cm. The gain of the cavity-backed spiral antenna is similar to the behavior of a dipole over a ground plane, being a maximum when the cavity depth is about a quarter wavelength (approximately 5 cm) and dropping off rapidly for depths greater than 3/8 wavelength. The antenna gain is approximately 6 dB in an optimum design, with a 3 dB beamwidth of approximately 60 degrees. In comparison with the turnstile antenna on a ground plane, the cavity-backed spiral antenna may be less desirable for array applications due to cost and complexity. However, it can be used as a low gain radiator for mechanical pointing or as a beam switching element. A typical cavity-backed spiral design is shown in Figure 4-37.

4.13.2 Design Features of Cavity-Backed Spirals

Like the conical spiral, the simplest cavity-backed spiral design is a two-arm spiral of copper foil printed on a dielectric plate and mounted on a cavity. The two arms are fed balanced at the center terminations from coax lines through the cavity back wall from a stripline hybrid. Also, like the conical spiral, use of 4, 6, or more spiral arms improves wide angle polarization circularity and permits simultaneous sum (Σ)

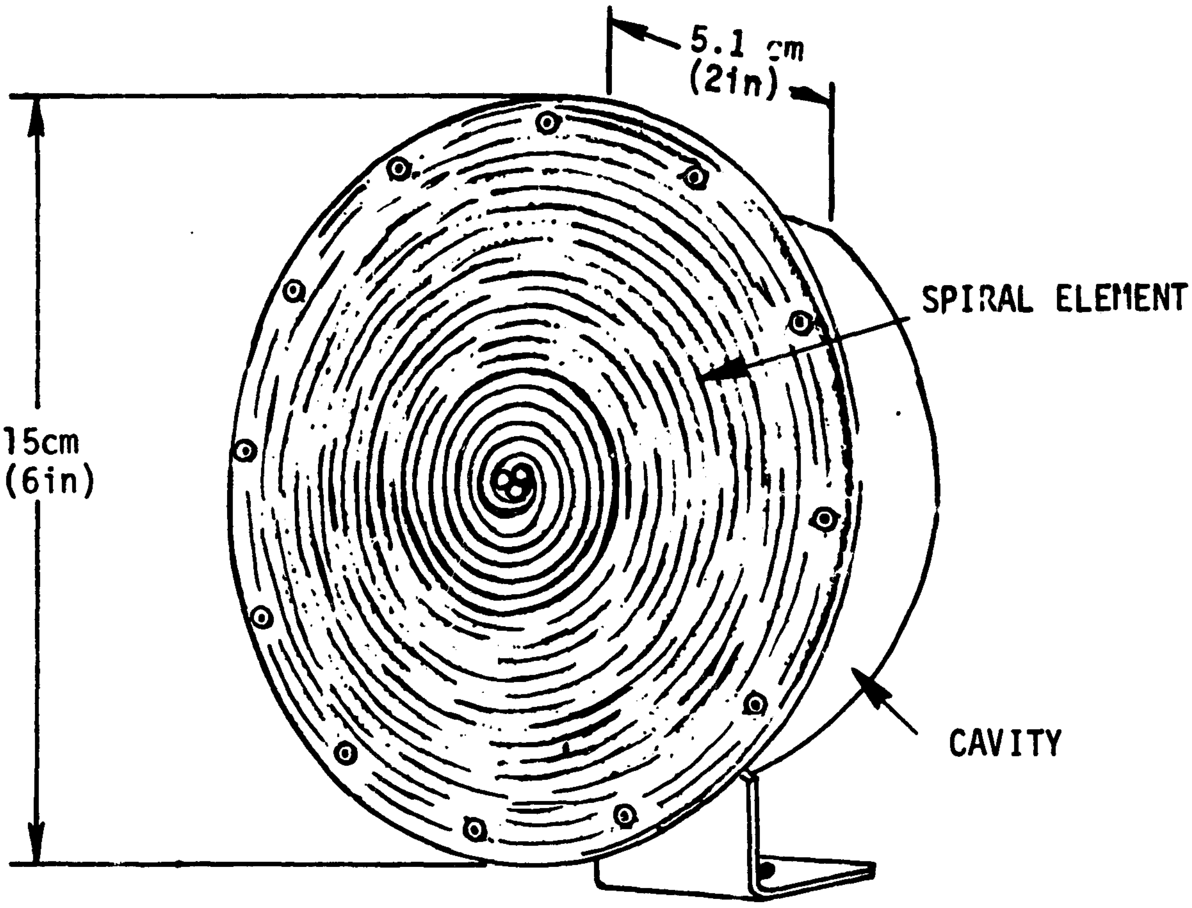

Figure 4-37 CAVITY-BACKED SPIRAL ANTENNA

and difference (Δ) patterns to be formed for dual channel monopulse feeding of a single antenna. Figure 4-38 shows the pattern performance of a 6-arm antenna.[52] The extremely low axial ratios over the wide pattern angles may be noted. However, 180° beamwidth axial ratios are inferior to those of an n-arm conical spiral. Both central and outer arm feeds have been used to provide independent circularly polarized monopulse patterns of both right and left hand circular polorization.

A disadvantage of the flat spiral is that the beamwidth is more or less fixed rather than being controllable by arm pitch angle as is the conical spiral. It is, however, quite small and compact and readily amen-

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(52) Chadwick, G.G. "Multiple Arm Spiral and its Derivatives For D/F and Homing" (Radiation Systems Inc.) Paper presented at Los Angeles Chapter IEEE, Group on Antennas and Propagation, 15 January 1970.

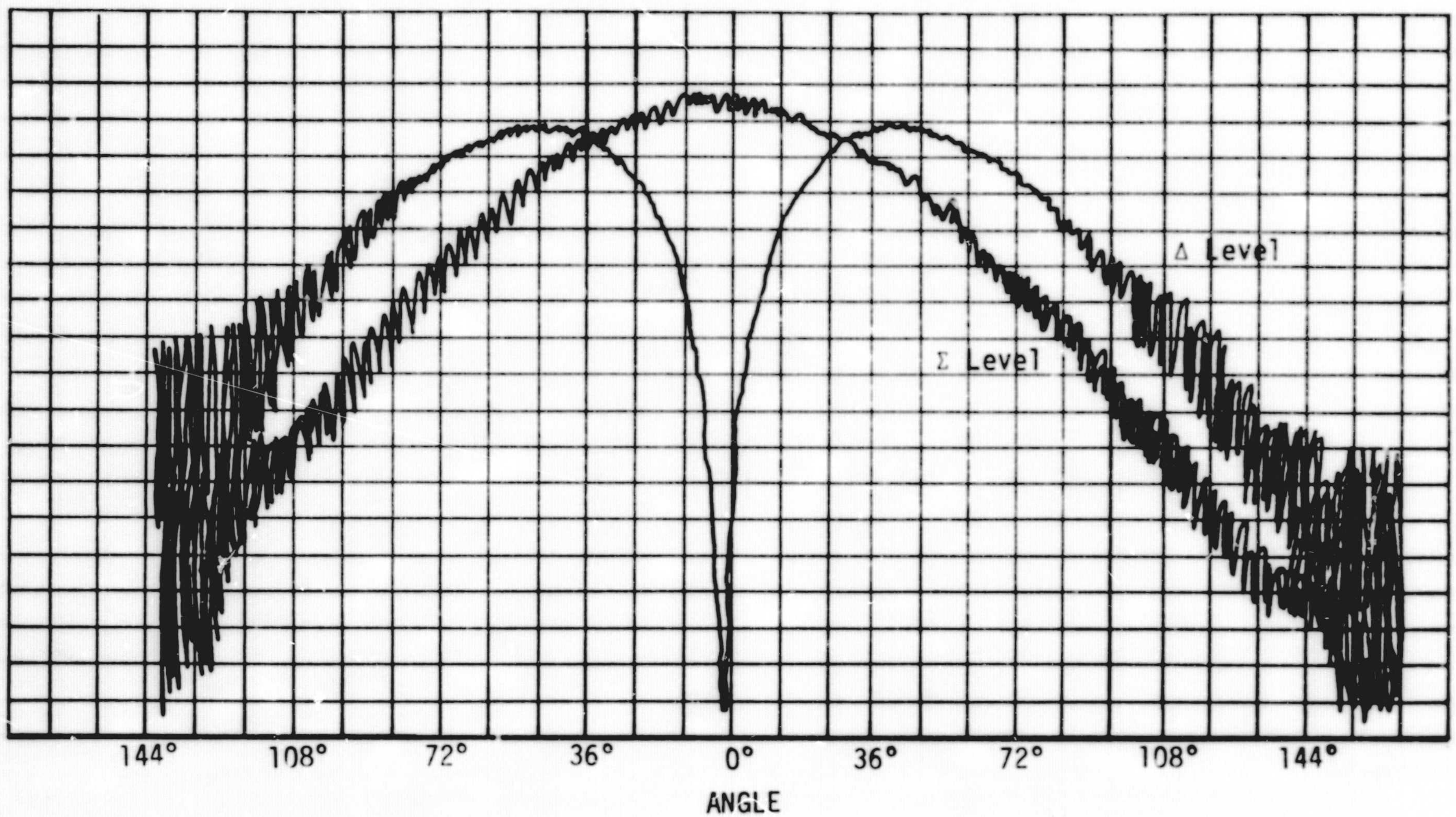

Figure 4-38 6-ARM CAVITY BACKED SPIRAL: TYPICAL Σ - Δ PATTERNS able to encapsulation in fiberglass, and should be considered a credible candidate for a gain requirement on the order of 6 dB.

4.14 CAVITY-BACKED CROSSED SLOTS

4.14.1 General Configuration

The crossed slot antenna is another design capable of producing a broad, circularly polarized pattern. The antenna consists of a pair of orthogonal slots backed by a cavity. A detailed arrangement of an L-band design is shown in Figure 4-39. In this design, each slot is symmetrically fed from the slot ends by means of coaxial lines. The necessary phase of quadrature feed for circular polarization is provided by a 90° hybrid as illustrated in the schematic diagram of Figure 4-40.

4.14.2 Performance Features

The peak gain of the antenna is approximately 4.5 dB above a circular polarized isotropic source, i.e., 4.5 dBi. The 3 dB beamwidth of the antenna is approximately 140 degrees. Axial ratios at 90° off the beam axis are typically 6 dB for crossed slots in a 0.5 wavelength diameter cylinder.[51]

Although the antenna has a desirably broad pattern for phased array application of large angle scanning, it is not very cost effective due to design complexity for either the mechanically pointed single-beam category or the scanned array category.

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[51] Scott, W.G. et al Op. Cit.

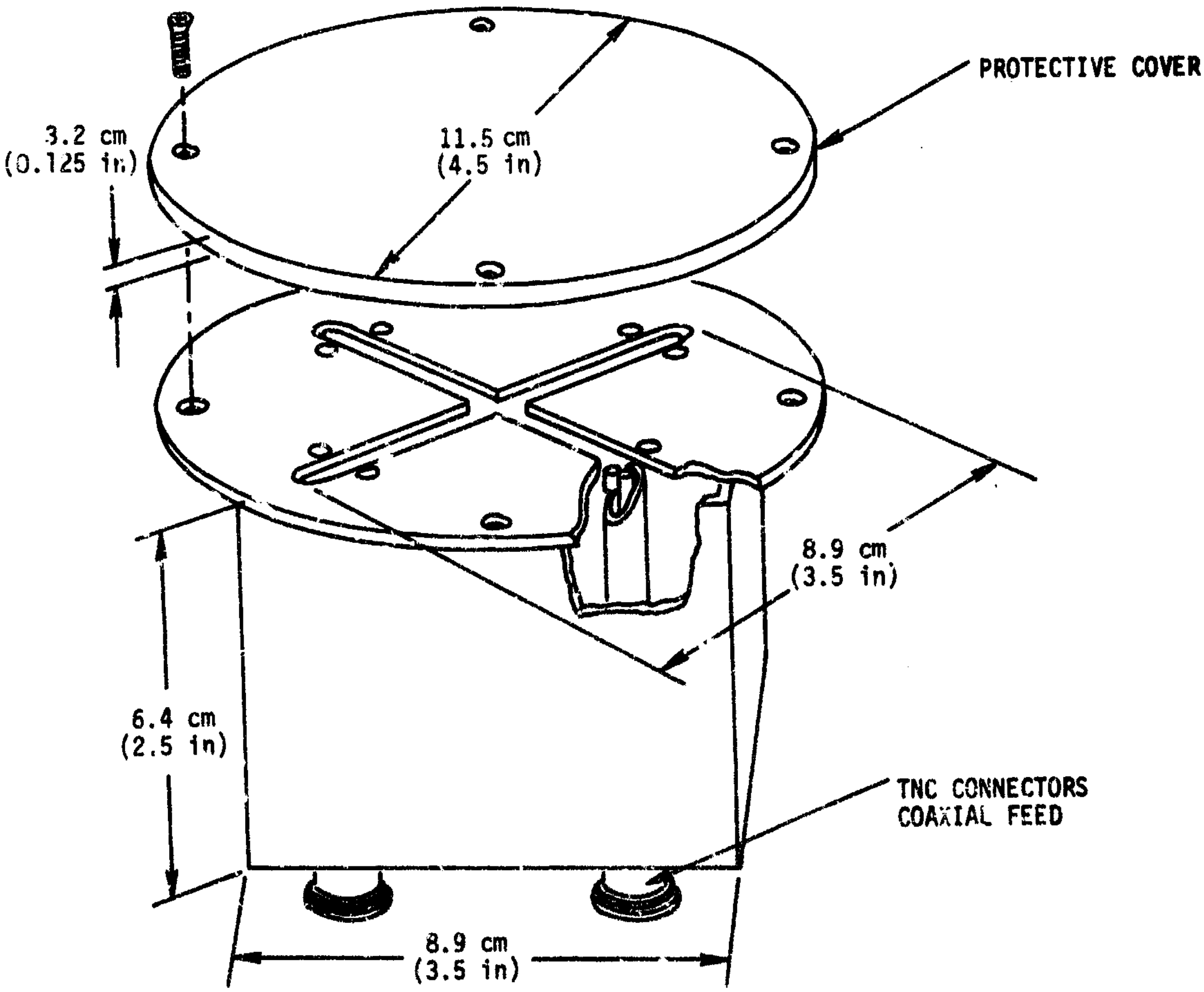

Figure 4-39 CROSSED-SLOT CAVITY BACKED ANTENNA

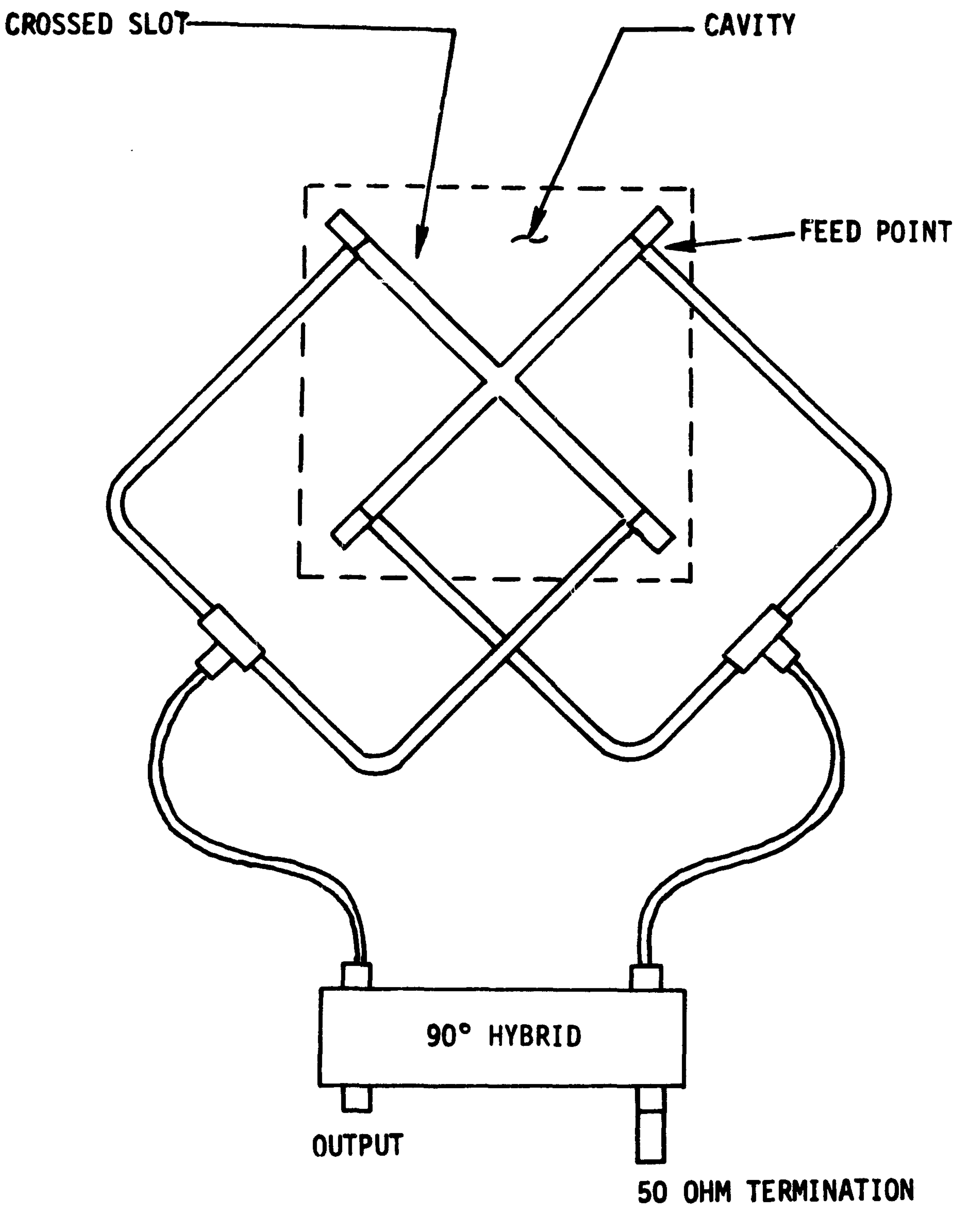

Figure 4-40 CROSSED SLOT ANTENNA ASSEMBLY SCHEMATIC DIAGRAM

4.15 CAVITY-BACKED CROSSED DIPOLES

This is an antenna design where a crossed dipole (turnstile) radiator is housed in a cavity. The antenna performs much like the cavity-backed crossed-slots antenna with the E and H fields interchanged. However, the general design is much simpler than the crossed slots since its excitation of the crossed dipole elements can be done by a simple split balun, similar to that of the turnstile antenna operating over a ground plane, described in Section 4.11. The beamwidth of the crossed dipoles is less than the crossed slot design. Its application is limited to low gain performance for mechanical pointing or switching design. A typical design is illustrated in Figure 4-41.

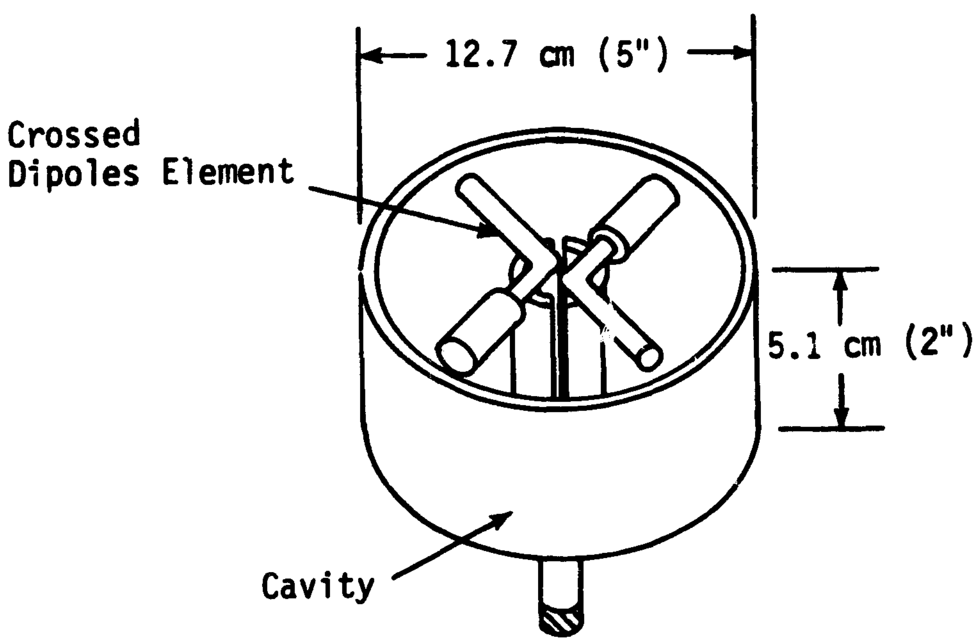

Figure 4-41 CAVITY BACKED CROSSED DIPOLES

This type of antenna is well suited to use in arrays because of its simplicity and small size. In such applications the cavity is very often made with a square cross-section to permit a simpler mechanical configuration for the array.

4.16 SUMMARY COMPARISON OF MECHANICALLY POINTED, SINGLE-BEAM ANTENNAS

A representative selection of antenna radiators which provide values of gain of between 3 dB and 18 dB have been studied in some detail. The performance constraints placed on the design of the antenna are quite modest, as is evidenced by the sheer number of candidate designs which would be capable of meeting the requirements. Physical sizes of all antennas at 1540 MHz are quite modest, especially for values of gain less than 15 dB, so that no particular emphasis on absolute (aperture) efficiency is required. In addition, severe problems of weight or wind torques do not exist to complicate pedestal design.

In general, the designs studied are quite broad-band. Even when optimized at 1540 MHz, the loss in efficiency at 1640 MHz is generally offset by operating at the higher frequency, so that gain values are comparable.

Other considerations of major importance are the capability of particular designs to produce circular polarization, to develop an axial null pattern for automatic angle tracking (preferably by single channel monopulse), moderate sidelobes and low complexity.

Estimates of cost are extremely difficult to make, since the type of antenna considered is not generally produced in quantity, and there is little industry experience in estimating costs for production quantities in the region of one thousand. It is evident, however, that the cost of the antenna radiating element is likely to be a small part of the overall terminal cost ($\sim$2 to 4%) so that large percentage variations in antenna radiator costs will have a very small effect on the overall terminal cost. The general characteristics of the antennas studied are summarized in Table 4-1 for comparison purposes.

TABLE 4-1 COMPARISON OF SINGLE-BEAM ANTENNA TYPES

PARAMETER / ANTENNA TYPE	GAIN (NOMINAL)	SIZE (AT NOMINAL GAIN)	BEAM SHAPE	BEAMWIDTH	GENERAL VOLUME	POLARIZATION	CAPABLE OF TRACKING?	SIDE AND BACKLOBES	COMPLEXITY, ETC. (RELATIVE COST*)
Parabola	15 dB (minimum)	30" (750 mm) min. Dia.	Pencil Beam	Typically 18° for 20 dB Gain	Moderately Large	Depends on Feed. C.P. readily achieved	Yes, with multi-element feed.	Low	Impractical at low gains (M)
Parabolic Cylinder	12 dB minimum	30"x5" (750x185 mm)	Fan Beam	Function of Aperture (17° x120° for 12 dB example)	Moderately Large	Depends on Feed. Linear readily achieved	Yes, in elevation, with multi-element feed	Moderate	Simple Fan Beam (M)
Planar Array	Gain varies with aperture size (8 - 18 dB)		Any Shape	Function of Aperture	Moderately Large	Depends on element. C.P. readily achieved.	Yes	Low	Versatile but Complex and Expensive (VH)
Linear Array	Gain varies with array length (9 dB ~ 780mm)		Fan Beam	Function of Array Length (120°x18° for 9 dB array)	Moderate but long	Depends on element. Can be made circular	Not Readily	Moderate	Less Complex than Planar Array (H)
Crossed Yagi-Uda	Gain varies with length (10-15 dB) ~ (400-1000 mm)		Wide Symmetrical Beam	30°-50° for 11-15 dB	Moderate but long	Circular	Yes, but only with linear polarization	Moderate	Light but fairly complex (M)

*RELATIVE COST: VH = Very High; H = High; M = Medium; L = Low.

TABLE 4-1 COMPARISON OF SINGLE-BEAM ANTENNA TYPES (Con't)

PARAMETER / ANTENNA TYPE	GAIN (NOMINAL)	SIZE (AT NOMINAL GAIN)	BEAM SHAPE	BEAMWIDTH	GENERAL VOLUME	POLARIZATION	CAPABLE OF TRACKING?	SIDE AND BACKLOBES	COMPLEXITY, ETC. (RELATIVE COST*)
Helix	Gain varies with length 9 dB∽8" (20 cm) 15 dB∽21" (54 cm)		Wide Symmetrical Beam	35° for 14 dB gain	Small	Circular	No	High	Small, Light, Simple to make (L)
Log-Periodic	9 dBi	38 cm x 51 cm (15"x20")	Wide Symmetrical Beam	50° at 9 dB	Moderately Large	Circular	Yes, but only with linear polarization	High	Limited to ∽9 dB gain (M)
Horn	14-19 dB	33-51 cm (12"to20")	Broad "Flat-topped" Fan Beam	Varies with gain (18° at 19 dB)	Moderately Large	Circular difficult with Sectoral Horns	Yes, with multimode horns	Moderate to high, depends on geometry	Fairly simple, but bulky (L)
Short Back-fire	14.5	40 cm (16")	Wide Symmetrical Beam	35°	Moderate	Circular	Yes, with multiple crossed dipole feeds	Low	Very efficient, Simple (L)
Turnstile on Ground Plane	3-6 dB	10 cm (3.94")	Very Wide Symmetrical Beam	80°-120°	Small	Circular	Yes, with multiple crossed dipole feeds	Moderate to Low	Very Small; Low Gain (L)

*RELATIVE COST: VH = Very High; H = High; M = Medium; L = Low.

TABLE 4-1 COMPARISON OF SINGLE-BEAM ANTENNA TYPES (Con't)

ANTENNA TYPE \ PARAMETER	GAIN (NOMINAL)	SIZE (AT NOMINAL GAIN)	BEAM SHAPE	BEAMWIDTH	GENERAL VOLUME	POLARIZATION	CAPABLE OF TRACKING	SIDE AND BACKLOBES	COMPLEXITY, ETC. (RELATIVE COST*)
Conical Log-Spiral	2-8 dB	23 cm (9")	Very Wide Symmetrical Beam	60°-180°	Small	Circular	Yes, if multiple arms are used	Low	Small; Low Gain (M)
Cavity-Backed Spiral	6 dB	15 cm (5.9")	Wide Symmetrical Beam	70°	Small	Circular	Yes, if multiple arms are used	Low	Small, Low Gain (M)
Cavity-Backed Crossed Slots	4.5 dB	11.5 cm (4.5")	Very Wide Symmetrical Beam	140°	Small	Circular	Not Practical	Moderate	Small, Wide Angle (M)
Cavity-Backed Crossed Dipoles	3-6 dB	13 cm (5.1")	Very Wide Symmetrical Beam	80°-120°	Small	Circular	Yes, with multiple crossed dipole feeds	Moderate	Small, Low Gain (L)

*RELATIVE COST: VH = Very High; H = High; M = Medium; L = Low.

General statements can be made about various types of antennas for producing directive beams. Directive beams can be obtained fairly simply at L-band frequencies primarily due to practical physical sizes. For high gain applications, the paraboloid reflector is a widely used microwave antenna since it is simple and inexpensive. Broad-side arrays can also be designed to provide similar performance. However, they find limited use in practice owing to the number of elements required to approximate the reflector antenna performance. In the medium gain range, the end-fire or Yagi element and helical radiator can be designed economically to provide 15 dB gain. Other radiators such as the horn, corner reflector and short-backfire antenna can be implemented to provide approximately the same gain. Spirals, turnstile antennas and crossed slots are the most promising low gain radiators. These antennas are also well suited for array application to provide narrow beam and high gain performance.

From the candidates discussed, no one antenna design emerges as an obvious choice, since other system considerations will establish the major characteristics required. It is possible, however, to select antenna types which appear most suitable for use in certain gain ranges and this is done in Table 4-2.

TABLE 4-2 PREFERRED ANTENNA TYPES

Gain Range	Antenna Type	Remarks
3- 6 dB	Turnstile on Ground Plane	Small and Simple, Curved for improved axial ratios
7-10 dB	Helix <u>or</u> Horn	Both inexpensive, Horn better for tracking
11-15 dB	Short Backfire	Very efficient, Simple, Fair tracking
16-18 dB	Parabola	Simple and inexpensive at moderate gains

The array type antennas are absent from this table because of their complexity and consequent cost for fixed-beam applications. They have other advantages which are discussed in Sections 5 and 6.

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SECTION 5

MULTI-BEAM ANTENNAS WITH BEAM SWITCHING

5.1 CATEGORICAL CONSIDERATIONS

5.1.1 Basic Beam Switching Concept

An obvious alternative to the use of motors and gearing to direct a narrow-beam antenna is to use many antennas, all fixed at appropriate orientations and locations on the ship so as to provide the required coverage. In order to actually realize the required gain, the antenna receiving the strongest satellite signal must be selected. Thus, the basic concept of multiple-antennas with beam-switching is to obtain the complete required azimuth coverage (180°) and elevation coverage (90° plus maximum ship roll angle) by switching between a number of fixed angular coverage beams.

Unlike the mechanically-steered, single beam approach, the antenna radiators may be fixed to the ship structure. Conceptually, the motors and cable-wraps (or rotary joints) are replaced by the use of many antenna radiators and switches. The switch control, i.e., beam pointing, may be done remotely; either by manual operation, slaving to shipboard inertial references, or by comparing outputs from adjacent beams and angle tracking automatically. A process referred to as beam interpolation may be used to minimize gain variations as the signal direction moves across sectors.

In addition to the deletion of the requirement for motors, gears and cable wraps or rotary joints, a major advantage of the switched-beam concept is that the radiating apertures for the individual sector beams may be

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separated and mounted in strategic locations about the ship, rather than in a single co-located array of radiators. The separated locations could, for example, be at positions around one of the bridge decks, thus avoiding a need for mast location.

However, separate locations imply the use of multiple low noise receiver preamplifiers and possibly transmitters. While the transmit link can tolerate some line loss, the receive path may not (depending on radiator gain), so that each radiator may be required to have at least one diplexer, preselector filter and low-noise preamplifier. The line losses due to distance and/or insertion of switches can be compensated by larger (higher gain) antennas, and thus more of them to provide the required greater-than-hemispherical coverage. For a given radiator size, a considerable amount of additional electronics is required to implement an array of switchable beams relative to a single mechanically-pointed beam. The cost of this additional electronics is the major disadvantage of the switched beam concept. A second disadvantage for the shipboard application is that the antenna "system" would be considerably larger, in combined size, than the swept volume of a single movable radiator.

The factors which constitute these advantages and disadvantages are discussed in the following sections.

5.1.2 The Switch Location Problem

A fundamental consideration in the design of switched beam antennas - whether separate antenna apertures or multi-beam shared single apertures are used - is that of the location of the beam switch. Figure 5-1 illustrates the basic concept of switching, showing a single switch between antenna ports.

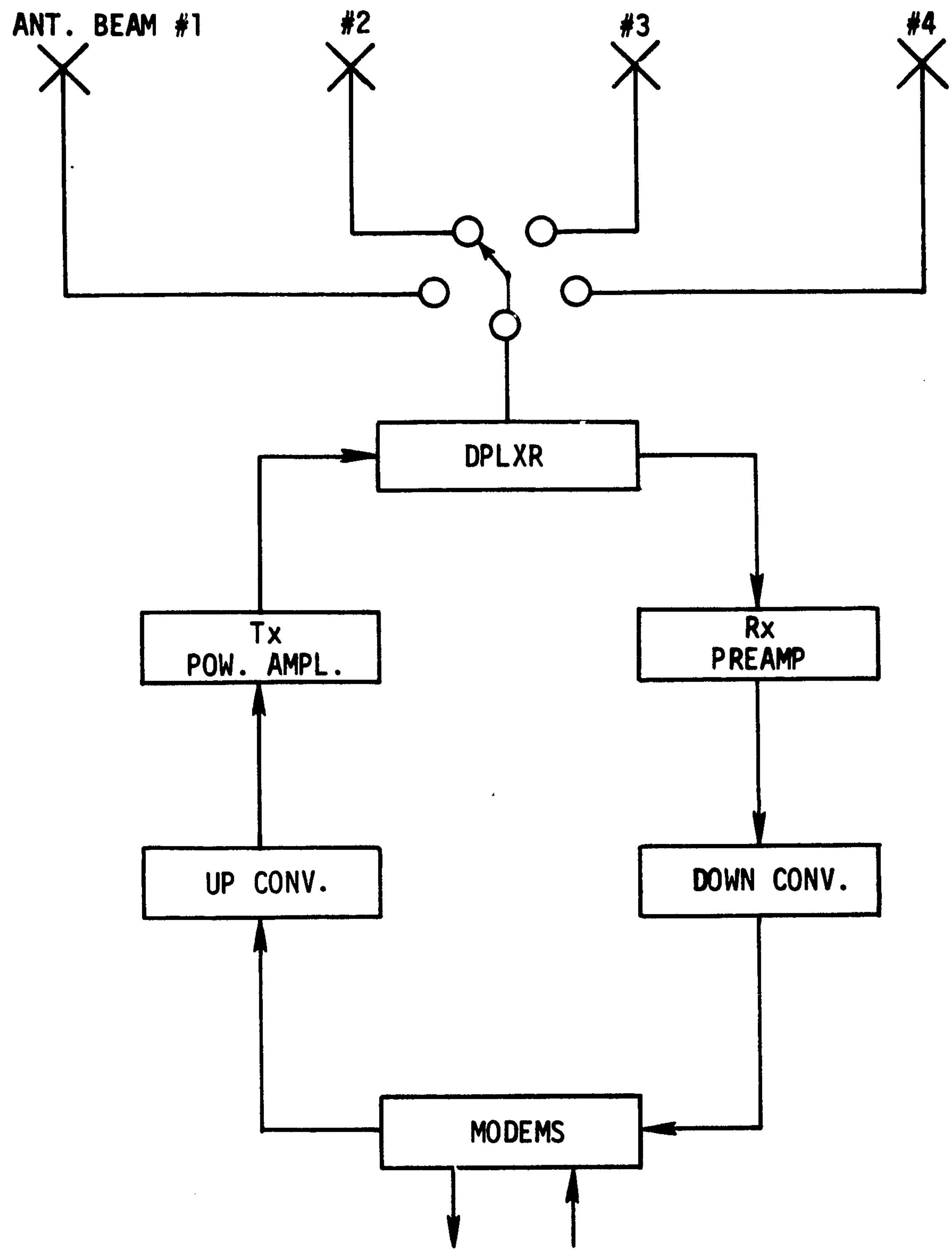

Figure 5-1 SWITCHED-BEAM CONCEPT WITH SINGLE ELECTRONICS GROUP

Clearly, if a single diplexer and transmitter/receiver are to be used, the separation between the antennas should be small. The line loss of typical coaxial cable for such an application, 1.3 cm (0.5 inch) in diameter, is greater than 4 dB per 10 meters (13 dB per 100 ft). More expensive coax or elliptical waveguide could be used to decrease this line loss, but such lines are not generally desirable for shipboard installation. With the beam switch located as in Figure 5-1, any line loss, including the loss of the switch itself (typically 0.5 to 1.0 dB) detracts directly from the antenna's receive and transmit gain. The bridge decks on large ships might be 30 by 30 meters. Large line losses such as 4-8 dB would be prohibitive, in terms of making them up by increasing the gain and number of antennas accordingly. Thus, if one common electronics group is to be used with switched-beams, the antennas should be virtually co-located, at least within 1-2 meters. Given this conclusion, it is appropriate to consider "shared" apertures using an array of elements such as described in Section 5.3.

The alternative to more and larger apertures due to line loss is multiple transmitters and receivers. This is conceptually illustrated in Figure 5-2. Use of multiple active components permits antennas to be separated by long distances. For even a nominal amount of net gain in such a concept, such as 9-10 dB, the number of antennas required can become prohibitively large.

5.1.3 Number of Beams for Switching

While it was estimated in Section 3.2 that the number of separate radiators (or beams) required to cover the hemisphere with a gain of G is about $2|G|$, the actual number is slightly more complicated, depending on beam shape, allowable cross-over nulls, etc. For example, a

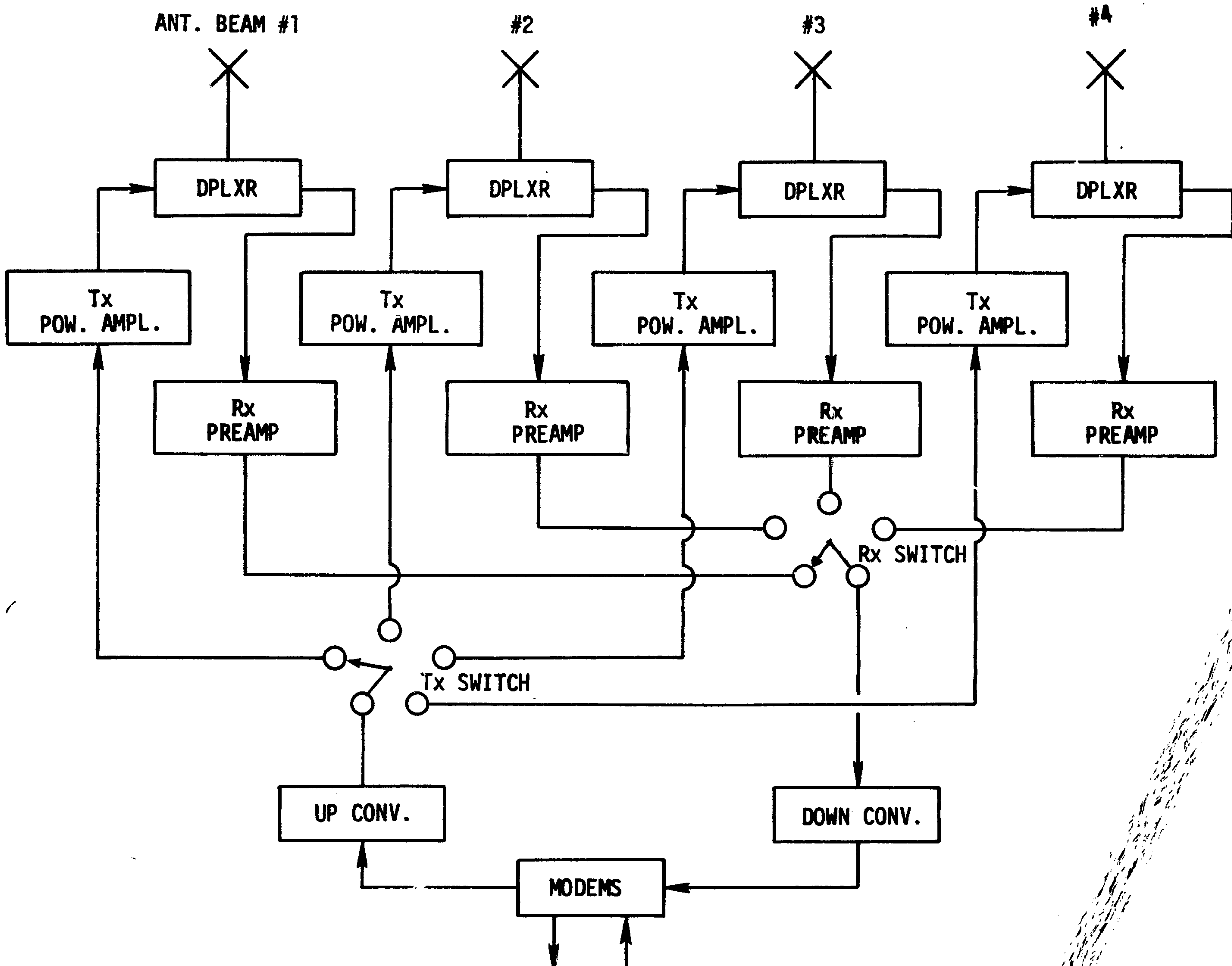

Figure 5-2 SWITCHED-BEAM CONCEPT WITH MULTIPLE TRANSMITTER/RECEIVERS

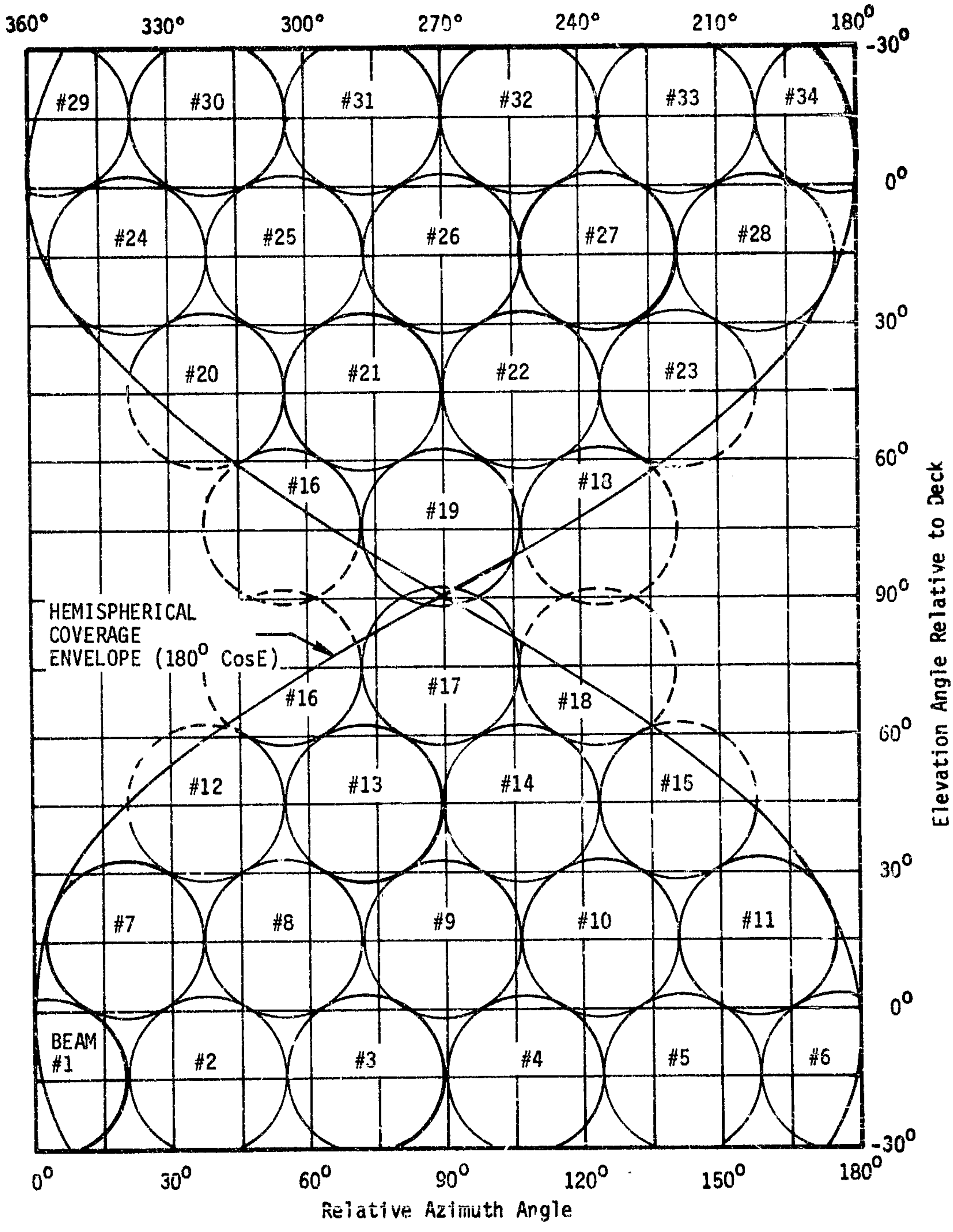

Figure 5-3 SWITCHED BEAM ANTENNA COVERAGE WITH 10 dB MINIMUM GAIN

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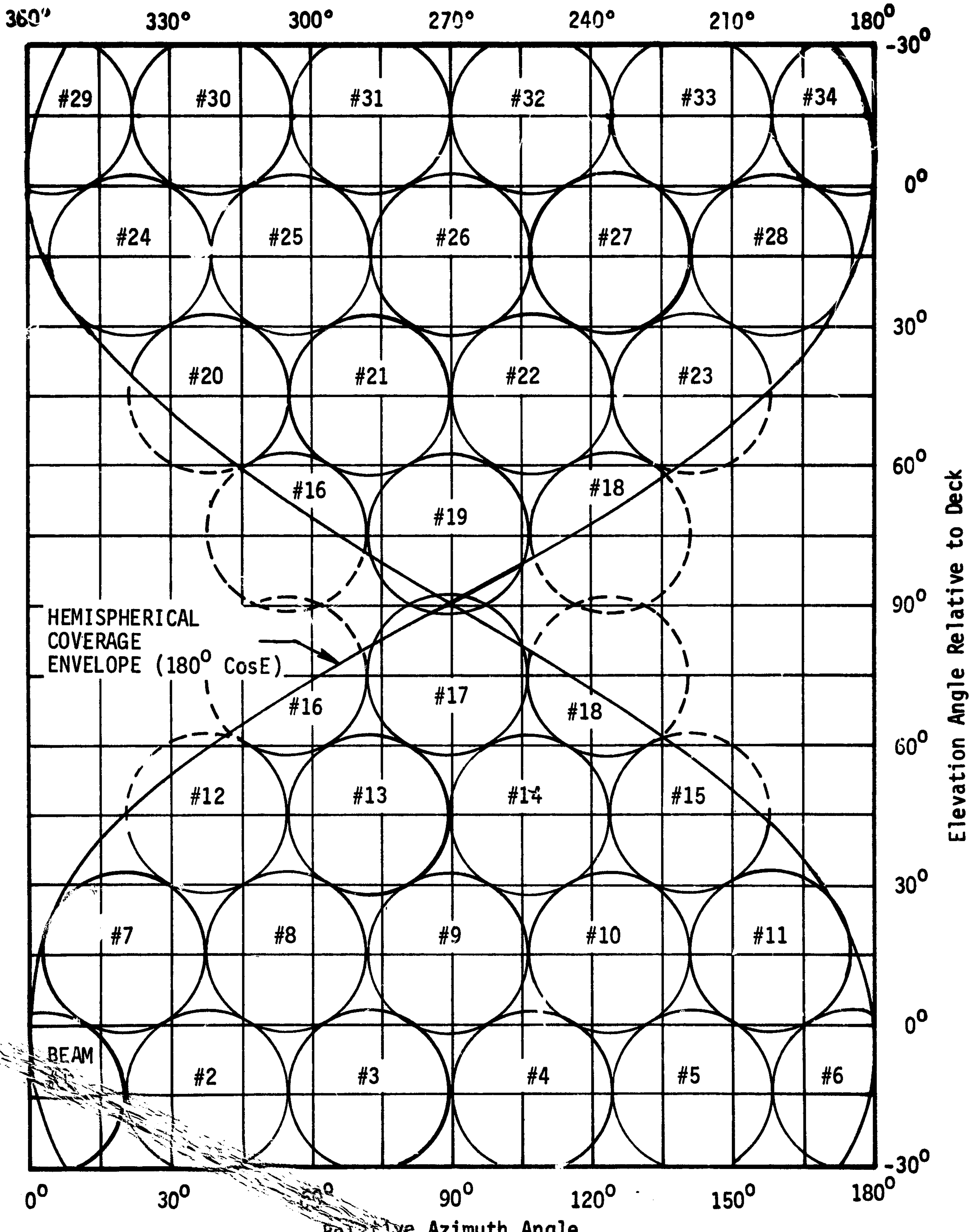

Figure 5-3 SWITCHED BEAM ANTENNA COVERAGE WITH 10 dB MINIMUM GAIN

receiver RF stages. In each location, one-half or one-quarter of the antenna radiators could be essentially attached to each other.

The short-backfire antenna was used as an example because of its small size (≈ 40 cm diameter by 10 cm). Other elements, such as rectangular horns, could also be used - which could yield less area below 3 dB points, but would require some form of circular polarizers, such as screens. In any case, the physical mass of 34 individual antennas would certainly be a significant burden to the ship. In addition, initial installation and alignment to set and orient each radiator individually would constitute a major cost disadvantage in such a system. Lowering the gain requirements obviously would lower the number of separate antennas required. For example, minimum net gains of 3-4 dB could be achieved with about 6 beams of 6-7 dB peak gain each.

While many configurations are possible, Table 5-1 below indicates the approximate number of separate beams required for selected values of minimum gain achieved.

TABLE 5-1

APPROXIMATE NUMBER OF FIXED BEAMS REQUIRED FOR SWITCHING

Minimum Net Gain Required	Typical Number of Beams
3-4 dB	6
6-7 dB	12
10-11 dB	34
15-16 dB	96

In order to reduce the physical mass involved in the above example, phased arrays can be used in which "groups" of beams are formed by single apertures containing many smaller elements. Such a concept is discussed in Section 5.3.

further general disadvantage associated with the switched beam category is the difficulty of automatic tracking. In theory, automatic tracking of the satellite signal is possible by comparing received levels in (four) adjacent beams. Sum and differences patterns can be established by appropriate comparator circuitry, which is switched from beam set to beam set, where a set consists of four or five beams. This would increase the number of RF switches in the receive path of Figures 5-1 or 5-2 by at least a factor of four.

In general, the increased complexity of automatic angle tracking with switched beam arrays tends to make the slaved stabilization approach (Section 3.4) more attractive in this antenna category. The circuitry required for automatic tracking is discussed in Section 5.3.3. As a major conclusion indicated above is that phased-arrays can be used to provide switched beams with smaller overall apertures, the following section reviews certain features associated with possible array elements.

5.2 ANTENNA ELEMENTS FOR SWITCHED-BEAM ARRAYS

5.2.1 General

In Section 4, a number of antenna types were discussed relative to their properties as individual radiators. In some cases, comments were made as to their suitability in arrays. The main criteria relative to suitability are beamwidth, i.e., coverage, and polarization axial ratio limitations when the elements are used in arrays.

Generally speaking, all the antennas discussed in Section 4 could be used in a switched-beam array configuration. Typically, low gain elements are most desirable in that they permit wide fan beams to be generated with a linear array. The fan beam approach with linear arrays is less complex than the planar array approach when considering beam switching. It simplifies the pointing circuitry, i.e., the switching logic used in slaving and/or the beam comparison circuitry used in automatic tracking. Of the elements discussed in Section 4, the following are especially good selections:

- Helices
- Turnstiles
- Conical log spirals
- Cavity-backed spirals

For some applications, i.e., the wide fan beam linear array, the above elements may have too much gain to yield the required wide angle, e.g., a -30° to 100° elevation angle in an array switched in azimuth only. This fan width is then about 130° for "large" ships (>150 meters). For "small" ships (<150 meters) with baseline worst-case roll amplitude of 45°, the elevation coverage requirement would be larger, i.e., approaching 145°.

Multi-arm conical spiral elements were examined in Section 4-5 for incorporation in a four element linear broadside array to produce a sectoral beam with wide angle circular polarization. In this section, several other elements are shown which may have structural, cost, or other advantages for particular installations in such arrays. These elements are: the resonant quadrifilar helix, the square waveguide septum horn and slot-dipole combinations. For comparison purposes, the off-axis performance of a standard half-wavelength turnstile is shown in Figure 5-4.

5.2.2 Resonant Quadrifilar Helix

The "resonant quadrifilar helix"(53)(54) is a short 4 arm helix turnstile driven at one end and shorted via crossed rods at the other end as shown in Figure 5-5. It is, in effect, a "wrapped-up" turnstile antenna which trades bandwidth and isotropic coverage (turnstile) for a cardioid of revolution, broad directive beam with very good wide angle circular polarization. Webster(55) has examined 1/4, 1/2, 3/4 and 1-turn structures with total resonant arm lengths of 1/2 wavelength. Also, similar antennas with $\lambda/4$ arms have been developed.(56) The best wide-angle axial ratio result is for a $\lambda/2$ arm, 3/4 turn version wherein axial ratios were under

(53) Kilgus, C.C. "The Resonant Quadrifilar Helix." IEEE Trans. Antennas and Propagation, May, 1969. (Correspondence)

(54) Kilgus, C.C. "Resonant Quadrifilar Helix Design." Microwave Journal. December 1970.

(55) Webster, C.W. "Test Report on the 3/4 Turn Resonant Quadrifilar Helix." Internal Memo S2T-4-037, October 10, 1968. Applied Physics Laboratory of the Johns Hopkins University, Silver Springs, Md.

(56) Webster, C.W. "Experimental Test Results - $\lambda/4$ Volute." Internal Memo S2T-4-042, May 1, 1969. Applied Physics Laboratory of the Johns Hopkins University, Silver Springs, Md.

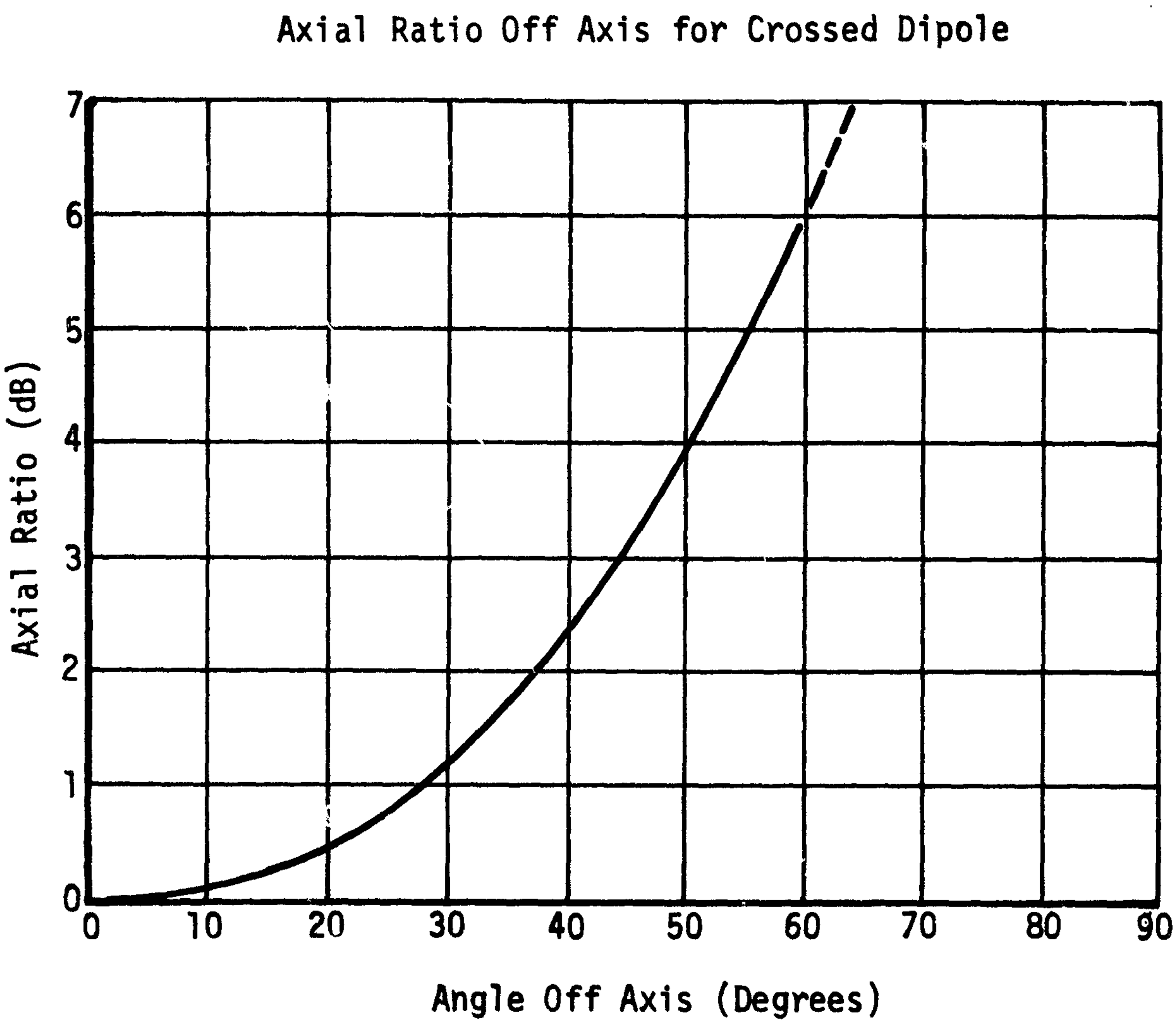

Figure 5-4 WIDE ANGLE AXIAL RATIOS OF HALF-WAVE TURNSTILE ANTENNA

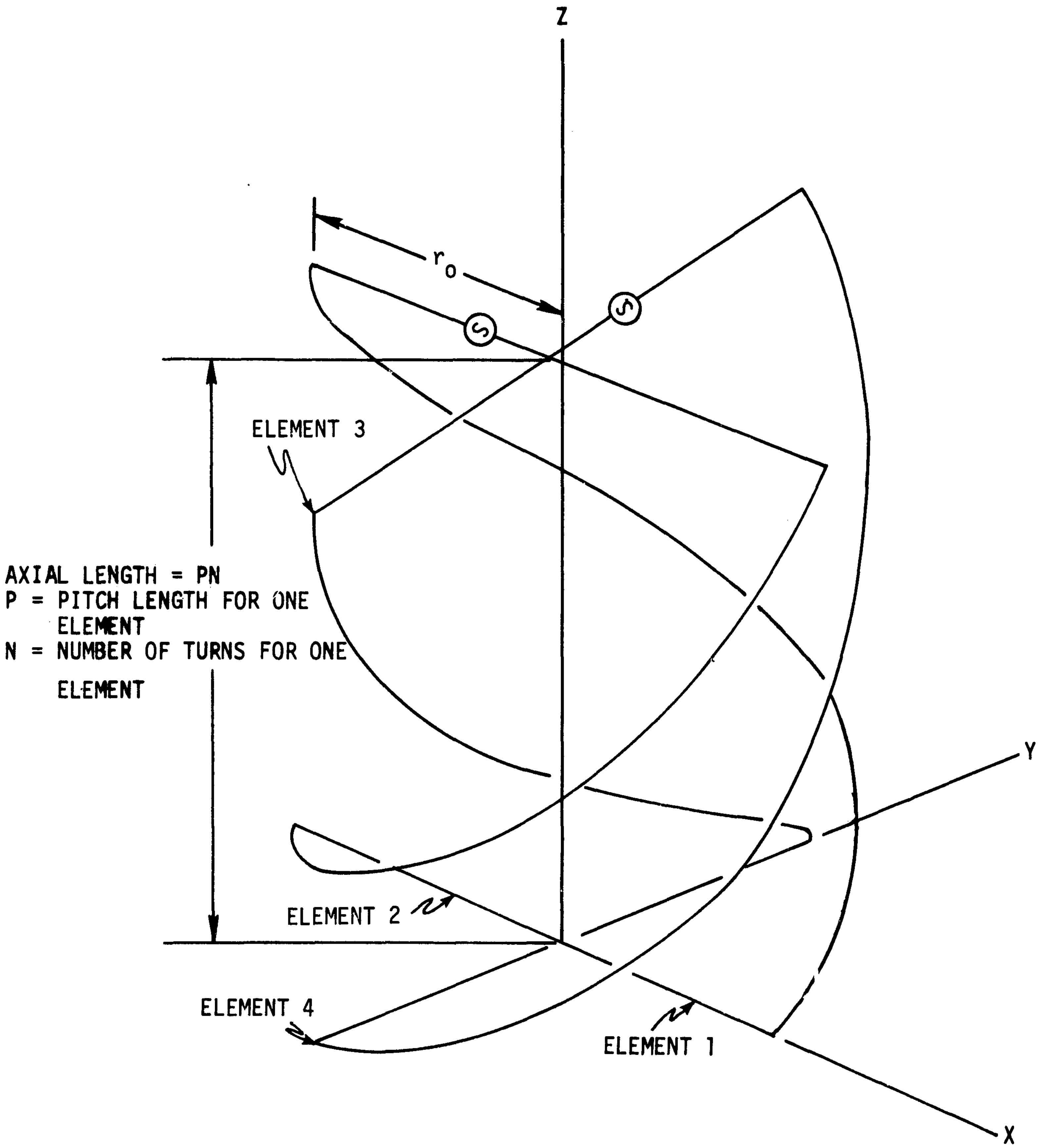

Figure 5-5 THE RESONANT QUADRIFILAR HELIX

3 dB for the entire main beam hemisphere. The half-power beamwidth was 110°, with amplitude about 9 dB down at 180° beamwidth. The maximum dimension would be 6.3 cm (2.5 in) (length) at L-band. The $\lambda/4$ arm antenna is even smaller and capable of the same performance, but actually could be too narrowband even for the maritime satellite application and more sensitive to environmental changes.

5.2.3 Square Waveguide Septum Horns

Another new element is the square waveguide septum horn,[57] shown in Figure 5-6. This antenna element is all metal and thus could be environmentally rugged. The circularly polarized horn shows exceptional axial ratios to wide angles in two orthogonal planes containing the aperture normal. For pattern planes between the two principal planes, axial ratios are comparable to those of a turnstile (Figure 5-4). However, this interplane performance is not detrimental for a sectoral beam with the narrow beamwidth under 20°, as is of interest in the subject application. A five element test array was tested and confirms this.[57] The wide fan, 10 dB beamwidth was 120°, with axial ratios under 2 dB. The narrow 3 dB beamwidth was 12°, with axial ratios under 2 dB through the first two (13 dB) sidelobes. However, at 90° off-axis axial ratios were 4 dB. The crossed-ellipse technique of Section 4.5.2 could perhaps improve these figures.

5.2.4 Slot Dipole Combinations

A third type of element is the slot-dipole combination, an example of which is shown in Figure 5-7. This element produces 120° beamwidth

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(57) Davis, D., Digiondomenico, O., and Kempic, J. "A New Type of Circularly Polarized Antenna Element."

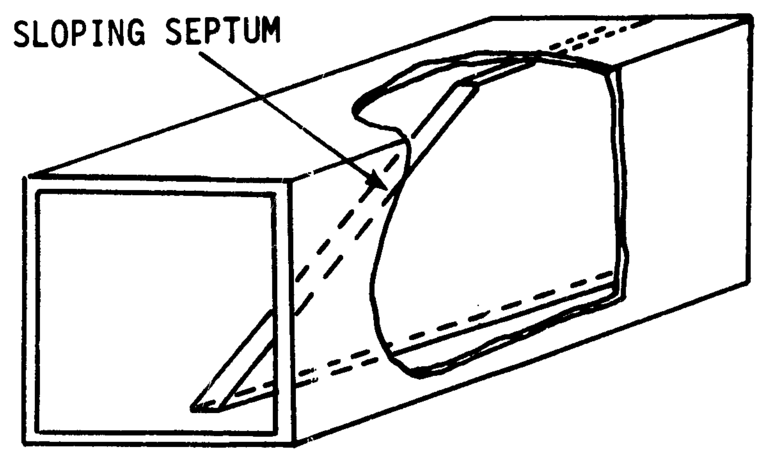

Figure 5-6   SEPTUM HORN

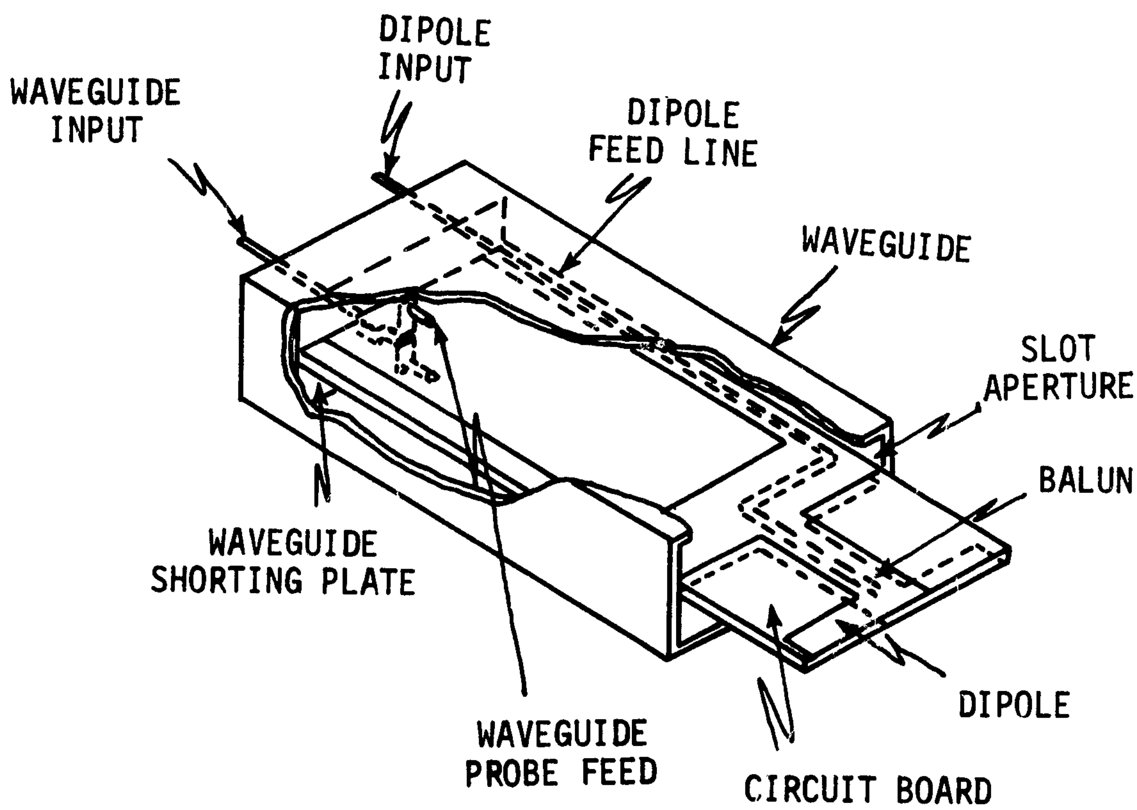

Figure 5-7 COMPLEMENTARY SLOT-DIPOLE ELEMENT

for axial ratios under 3 dB in its broader plane (-4 dB power points) and was developed specifically to minimize axial ratio degradation due to mutual coupling in the large array environment.[58][59] Its wide-angle performance is considered quite good, however its construction is fairly complex.

Thus, a considerably large variety of elements is generally available for use in arrays. However, wide angle coverage (for fan-beam linear arrays) is difficult to achieve. Because beamwidths of at least 180° would be required for azimuth fans which were switched in elevation angle only, elevation fans are considered simpler for the shipboard application in terms of wide-angle requirements, where the elevation beamwidth required is between 120° and 145°, depending on ship category.

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(58) Gabriel, W.F. and Dod, L.R. "A Complementary Slot-Dipole Antenna for Hemispherical Coverage." NASA GSFC Report X-525-66-435, TMX-55681, October, 1966.

(59) Cox, R.M. and Rupp, W.E. "Circularly Polarized Phased Array Antenna Element." IEEE Tran. Antennas and Propagation. (Correspondence) November 1970. pp. 804-907.

5.3 THE BUTLER ARRAY

The Butler array is a special type of switched-beam array employing an RF beam-forming device utilizing a matrix of 3 dB hybrid junctions and fixed phase shifters to form, in a theoretically lossless way, overlapping beams from a single array aperture. These overlapping beams are positioned in such a way that for linear arrays, the cross-over points are approximately 4 dB below their beam peaks. For a planar array, beam cross-overs are at -8 dB in diagonal planes. Each antenna beam has the gain of the entire aperture (allowing for the usual aperture scan loss). There are as many beams as the number of elements in the array, and the number of elements must be equal to a power of 2, i.e., 2, 4, 8, 16, etc.

All of the beam peaks and nulls occur at certain angles which are related to the array spacing by the following relationship:[60]

$$\begin{aligned} \sin\alpha_p &= (\lambda/Nd)\,(Nq + n - 1/2) \\ \sin\alpha_o &= (\lambda/Nd)\,(m + n - 1/2) \end{aligned} \qquad (5\text{-}1)$$

where α_p = angles of peaks

α_o = angles of nulls

N = number of antenna elements

d = element spacing

λ = wavelength

n = beam number ($n = 1$ for the first beam to the right or left of broadside, etc.)

q = any integer. ($q = 0$ yields the position of the main beam. Non-zero values of q give the position of grating lobes)

(60) Massachusetts Institute of Technology, Lincoln Laboratories. "Phased Array Radar Studies." Technical Report No. 236, Nov. 1961.

$m = 1, 2, 3$, etc. but $m \neq qN$

The angular coverage of the overlapping beams is:

$$\Theta_{pp} = 2 \sin^{-1} [\lambda/2d (1 - 2/N)], \tag{5-2}$$

where

Θ_{pp} = the angle between the peaks of the extreme right and extreme left beams

It may be seen from the above equation that for large angles of coverage, an element spacing close to $\lambda/2$ should be used. More important, the element spacing should be selected consistent with the capabilities of the array element coverage. For example, an eight element half-wave spaced linear array gives $\Theta_{pp} = 122°$ while for 16 elements Θ_{pp} would be 140°. To provide a beam pointing to a direction which has low element coverage results in a great reduction in gain. In designing an array, the element spacing and the element pattern must be carefully weighted. In addition to the gain reduction due to element pattern, the gain of the extreme beams is further reduced by the scanning angle. The beam broadens as it points away from broadside due to aperture foreshortening. For a high gain beam, a large number of elements must be used and the elements themselves must have a broad pattern.

The Butler array with a large number of elements is somewhat complex in the beam-forming matrix, where many hybrid junctions and phase shifters are required. For example, 12 hybrids and 8 fixed phase shifters are required for an eight element (eight beam) matrix. A 16 element (16 beam) array matrix requires 32 hybrids and 24 fixed phase shifters. However, this matrix is most compactly fabricated using strip-(transmission) line printed circuit techniques. Figure 5-8 illustrates an

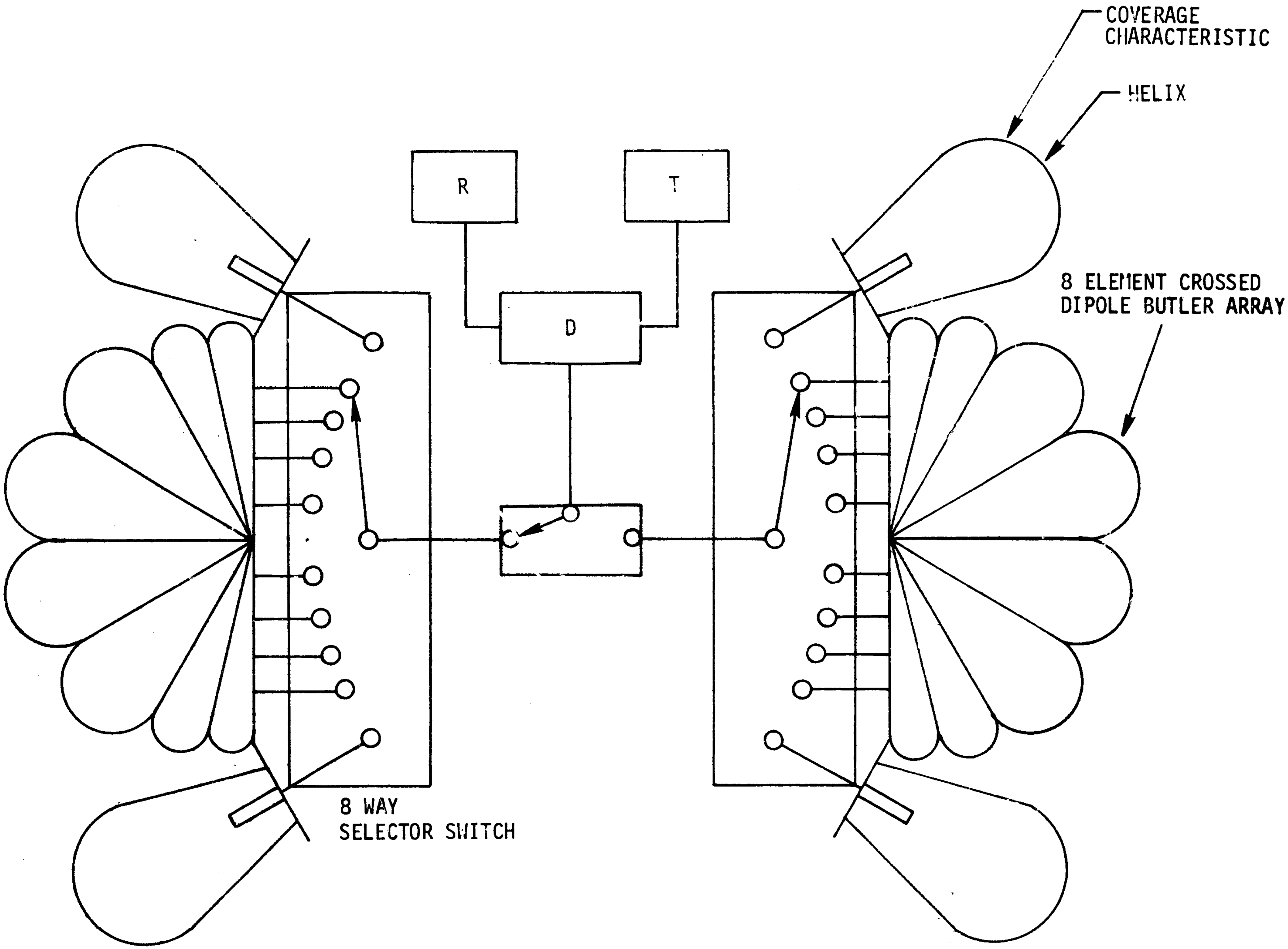

FIGURE 5-8 DUAL ARRAY SYSTEM FOR 6 dB MINIMUM GAIN OVER HEMISPHERE

eight-element linear array concept, in which the array is augmented by two helices to provide 10 overlapping beams.

Figure 5-9 shows the Butler array feed concept. Since each array and helix pair gives only half-hemisphere coverage, two such sets are required. The port and starboard sides of an upper bridge deck might be good locations for such an array. The basic beam concept is that of a fan beam, with a single wide angle beamwidth in elevation. The antenna sets should be tilted, at about 30° to 35° in elevation, about which the coverage is approximately ±60°. The peak gain of the array beam is approximately 10 dBi and at beam cross-over points, it is 6 dBi. At the expense of additional circuit complexity (available solid state power dividers and double-pole 8-way switches) the minimum cross-over gain can be increased to 9 dBi by beam interpolation.

As noted in Section 5.1, (and also in Section 6.4), either or both low noise receiver preamplifiers and transmitter power amplifiers may be inserted into the matrix. Also, by adding circulators and a second matrix, both transmit and receive operation may be accomplished from one array, using both pre- and post-amplifiers at each element, to avoid the line loss implied in Figure 5-9.

The minimum gain of +6 dBi for the dual Butler array system of Figures 5-8 and 5-9 is created by two factors. First, the individual multi-beam cross-overs drop 4 dB to this level. Secondly, the "hole filler" helix antennas drop to this level at their cross-overs with the fan beam array scan limits. The latter situation can be improved by using four Butler arrays of eight beams each configured as in Figure 5-10, with each array pointing at a 30-35° elevation angle but subtending only 90° in azimuth with its eight beams. With this configuration, the special "hole filler" beams are not required. However, the inherent

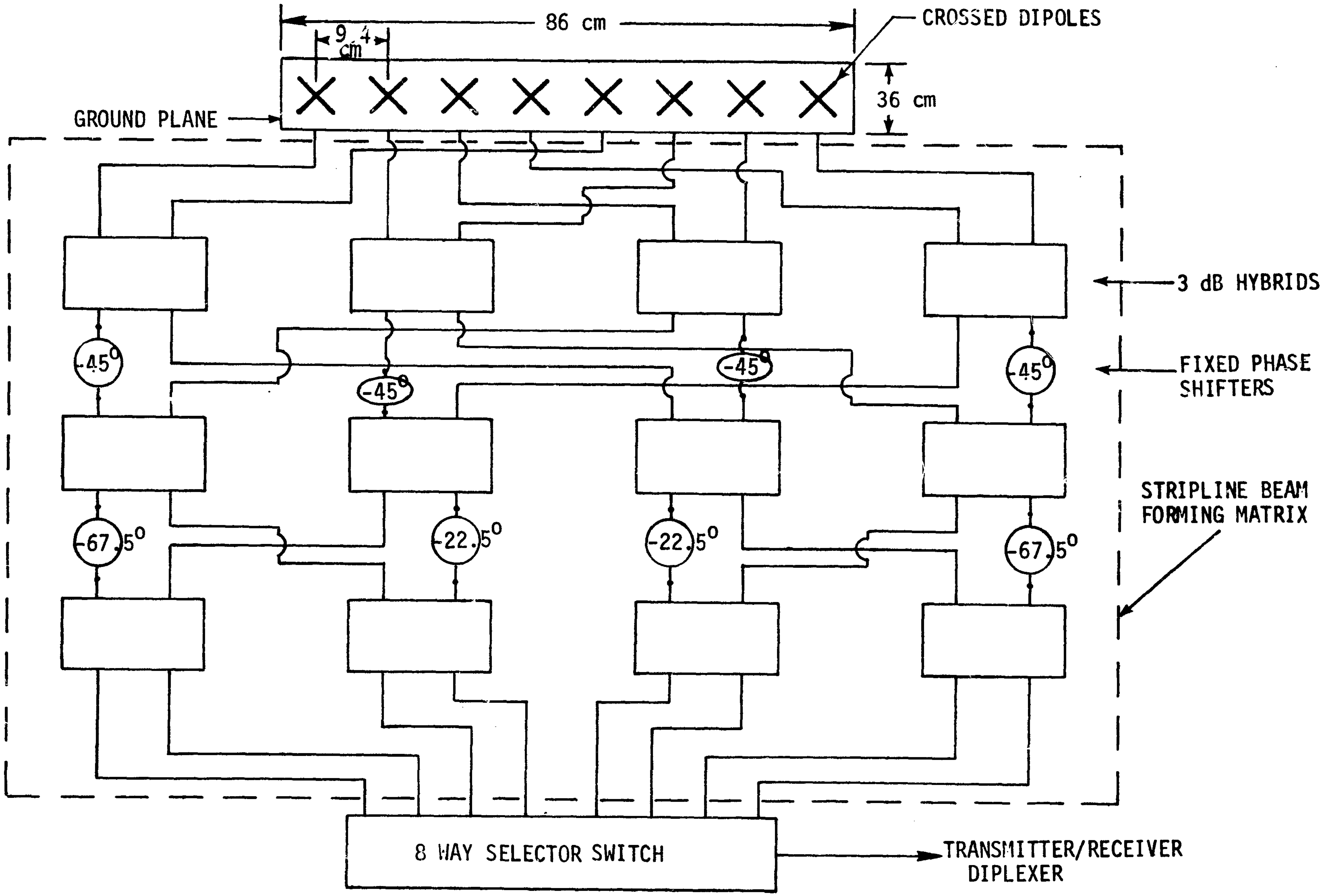

Figure 5-9 EIGHT ELEMENT CROSSED DIPOLE BUTLER ARRAY

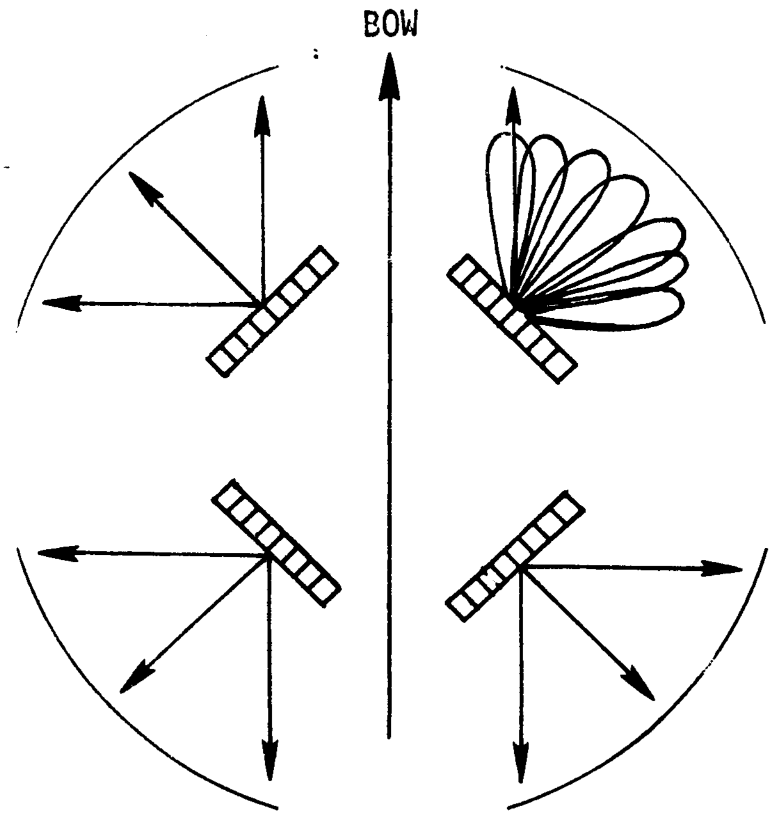

Figure 5-10 FOUR ARRAY SYSTEM FOR 10 dBi GAIN

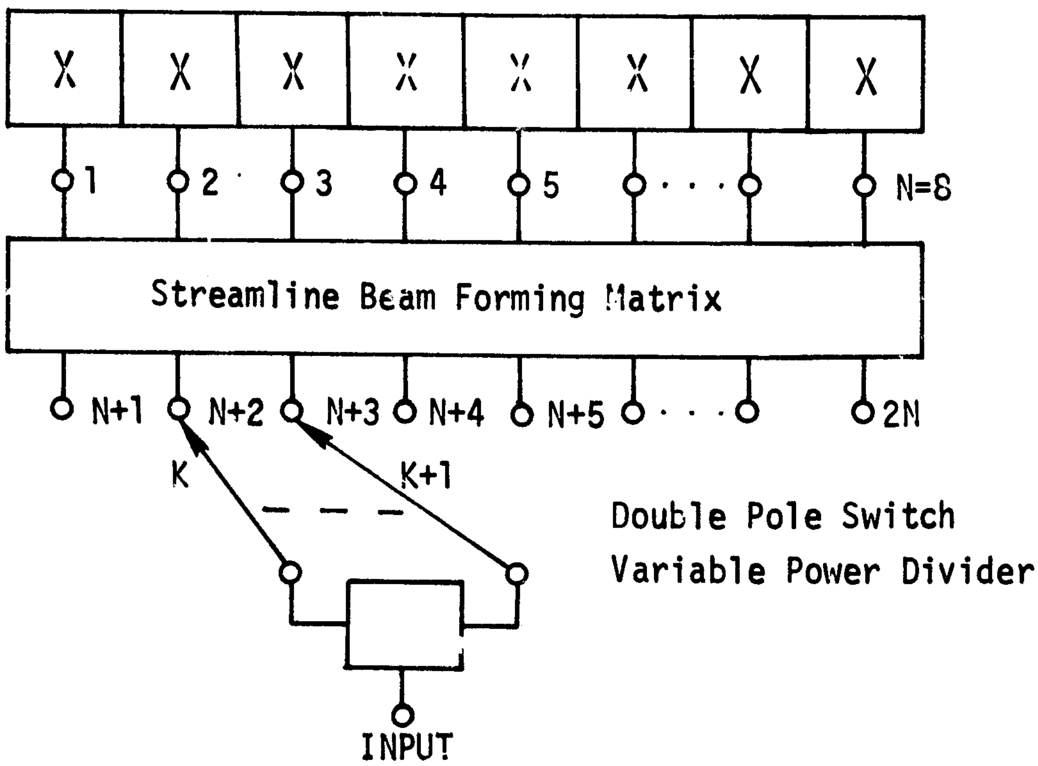

Figure 5-11 NULL-FILLING BY ADJACENT BEAM INTERPOLATION

6 dBi beam cross-over level is still present. But it can be improved, in this case, by additional RF circuitry to provide "adjacent beam interpolation", as shown in Figure 5-11.

The variable power divider permits the signal to be put into the "nth" beam, divided equally between the n and (n + 1) beams, or finally all into the (n + 1) beam. In the intermediate position, the composite beam has a 35 percent[61] wider half-power beamwidth than the individual beams. The composite beam falls at the cross-over of the n and (n + 1) beams and thus raises the effective cross-over level from 4 dB down to approximately 1.6 dB down, a considerable improvement. The variable power divider may be formed from solid state varactor or PIN diode stripline components. The divider may also be capable of continuous variation instead of having the three discrete positions described above. In this case, the beam interpolation provides continuous amplitude scanning across the angle between the n and (n + 1) beam positions and the gain ripple is only 1.3 dB below the n and (n + 1) beam peak gains. After the variable power divider has covered the n, (n + 1) sector, the double pole switch is stepped to the (n + 1) and (n + 2) terminals and then the beam interpolation is repeated.

Continuous monopulse null tracking can also be accomplished from the Butler array by the addition of more components referred to as a "steering box".[61] A block diagram of this automatic tracking array system is shown in Figure 5-12. It is basically a 3-way power divider using phase shifters as control elements. (Note that the phase shifters may be ganged to one control signal.) The summation of three adjacent beams permits the difference pattern to be generated for tracking in one axis. The sum beam peak and the difference beam null then track

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(61) Ross, G. and Schwartzman, L. "Continuous Beam Steering and Null-Tracking with a Fixed Multiple-Beam Antenna Array System." IEEE Trans. Antennas and Propagation. Sept. 1964. pp. 541-551.

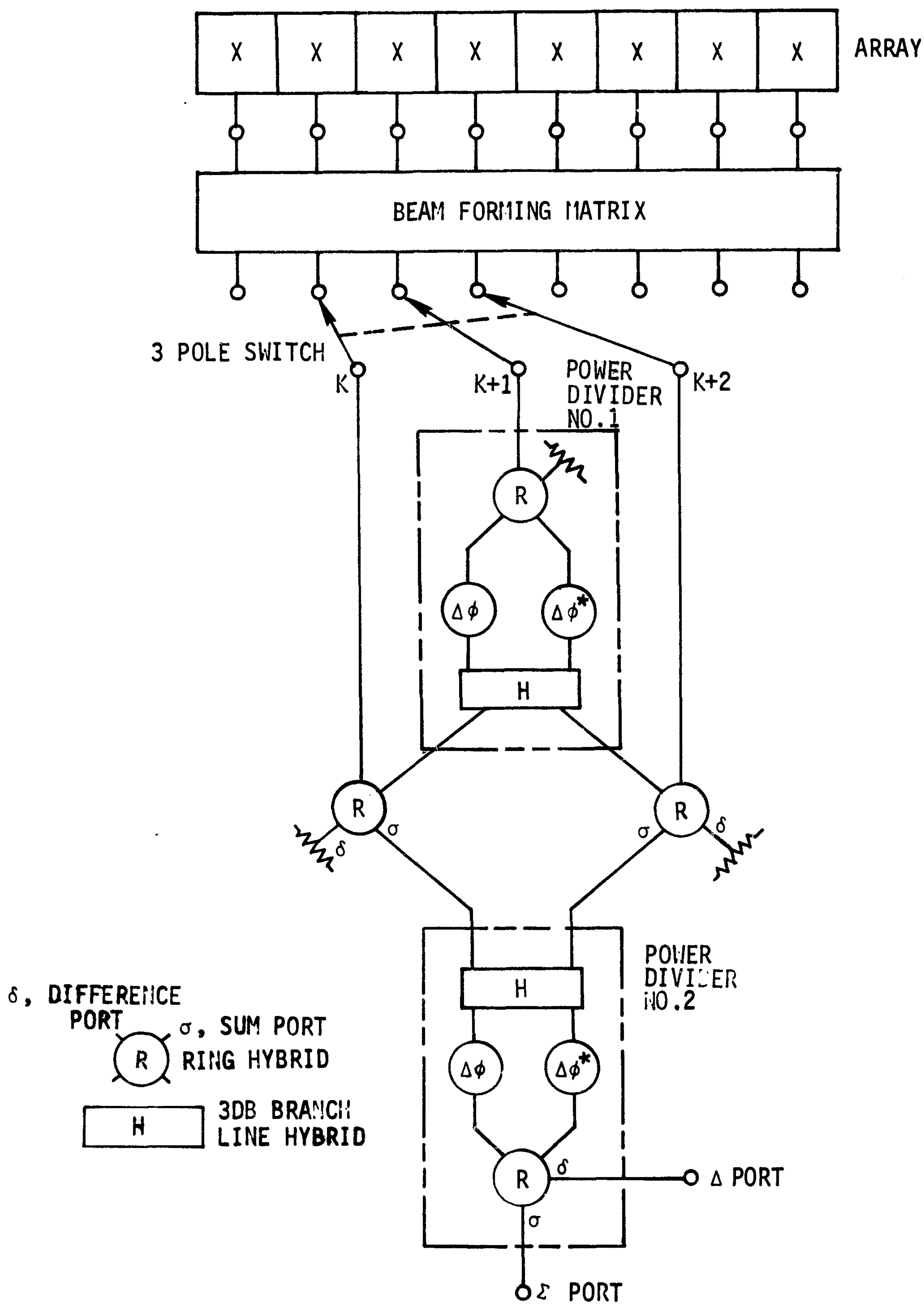

Figure 5-12 "LOSSLESS STEERING BOX" FOR LIMITED ANGLE TRACKING over the n to (n + 2) beam sector. A closed loop tracking system of this type is illustrated in Figure 5-13. Ross and Schwartzmann[61] show that the maximum change in the error sensitivity of the null pattern (i.e., the derivative of the difference pattern at the point of null signal) is $8/\pi$. Thus, the gain and phase margins for a closed-loop system must accommodate a change in error sensitivity of $8/\pi$, or about 8 dB. This could be a significant handicap, so that a "shaping network" might be necessary to maintain a constant scale factor.

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[61] Ross, G. and Schwartzmann, L. Op. Cit.

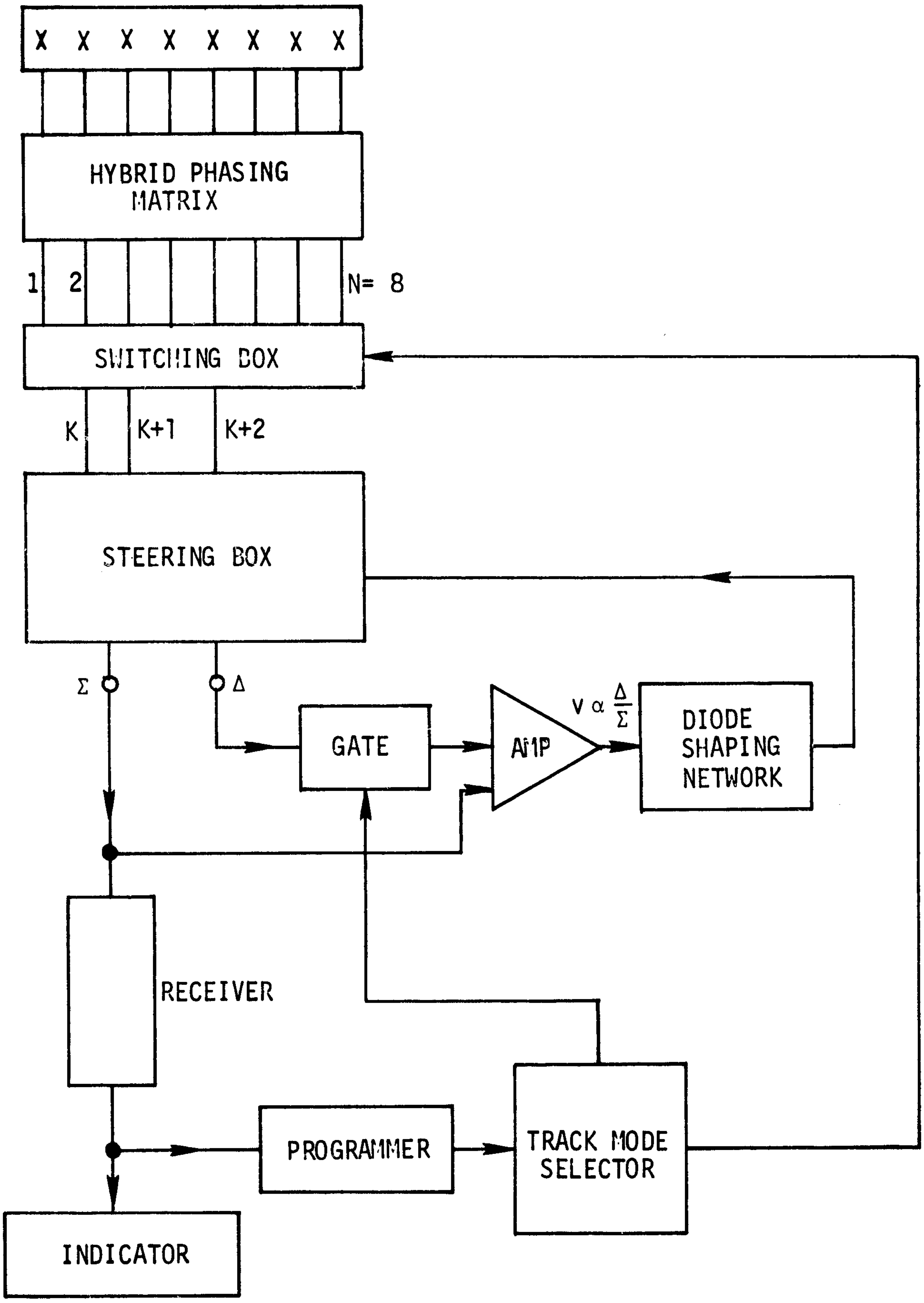

Figure 5-13 BUTLER ARRAY MONOPULSE TRACKING SYSTEM

5.4 ARRAY OF CORNER REFLECTORS

As one example of a switched-beam antenna which is not a shared array nor strictly separate independent antennas, an array of corner reflectors has been postulated which provides a peak gain of about 10 dBi. Figure 5-14 illustrates the basic concept.

Each of the six corner reflector elements provides a coverage of about 30° by 120°, and has a gain of about 9-10 dB. Without adjacent beam interpolation, the minimum gain of the antenna could be as low as 6-7 dB. The inherent polarization characteristic is linear along the half-wavelength dipole feeding the reflectors. With the aid of a wave-polarizing converter screen, the linear polarization may be converted to circular. The polarizer is located at the aperture of the corner reflector array, as indicated in Figure 5-14. The entire antenna array is tilted 30 degrees in elevation. The array measures about 2.6 meters in arc length. To gain full hemispherical view, two arrays must be used. The preferable locations for the antennas are on the port and starboard sides of one of the upper bridge decks, or of course, the mast. The simplest method for selecting antenna beams is via remote switching with reference signals from the ship's gyrocompass, i.e., slaving, although automatic angle tracking is possible.

The major disadvantages of this design are its large physical size and the cost of the polarizer design. An alternative technique is to rotate the feed dipole so as to tilt the dipole axis at an angle relative to the reflector corner. Woodward[62] has shown that the tilted dipole feed will cause circular polarized radiation from the corner reflector. This technique is somewhat cheaper to fabricate than the wave polarization converter but is probably somewhat inferior in off-axis axial ratio performance.

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(62) Woodward, O.M., "A Circularly-Polarized Corner Reflector Antenna," IEEE Transactions on Antennas and Propagation, July 1957, pg. 290.

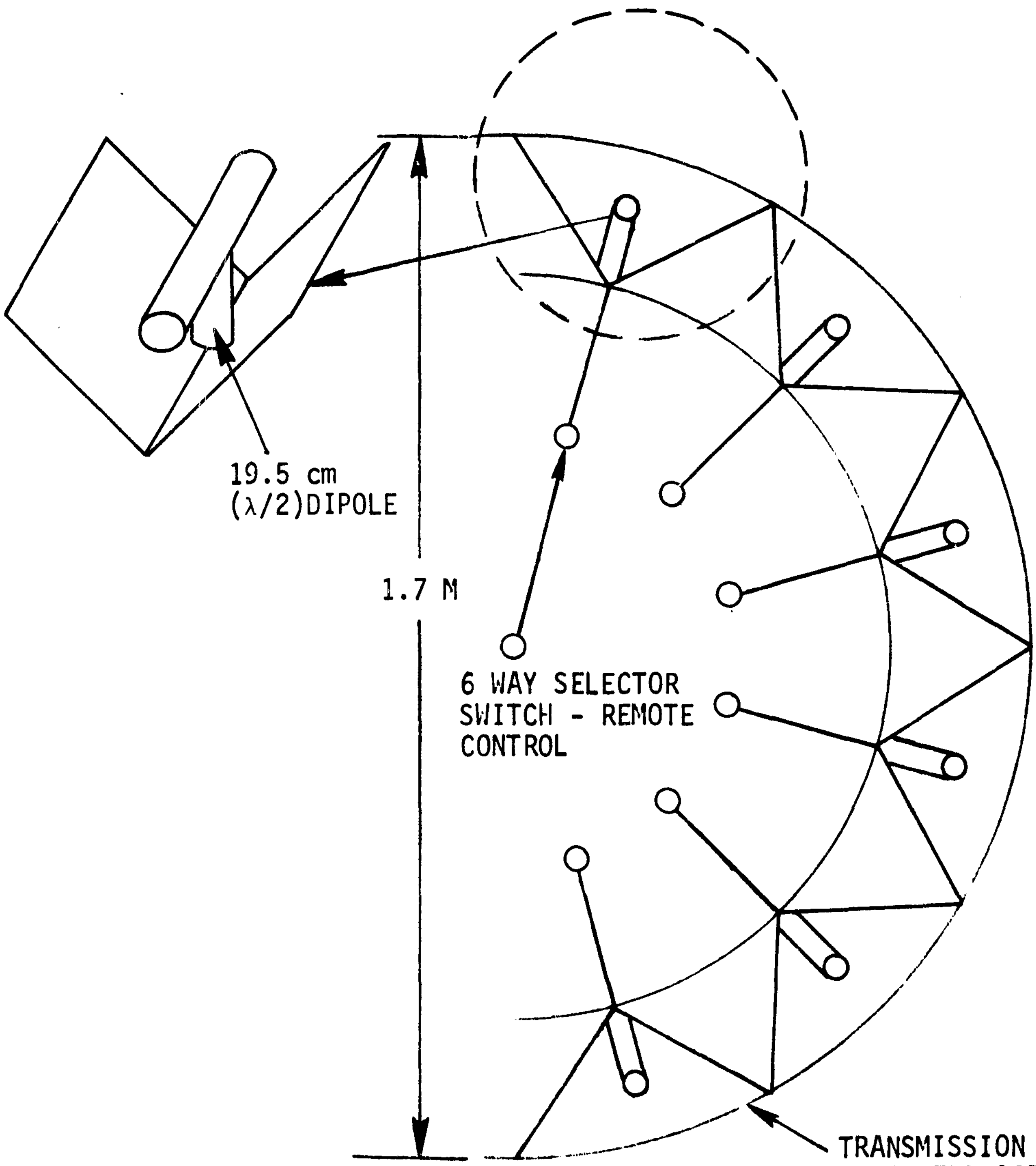

FIGURE 5-14 CORNER REFLECTOR ANTENNA FOR BEAM SWITCHING AND 9-10 dB PEAK GAIN

5.5 SWITCHING OF SEPARATE, FIXED ANTENNAS

In Section 5.1, it was noted that the beam switching concept can be implemented with separately located antennas, oriented to provide the necessary hemispherical coverage. An example was illustrated in Figure 5-3 for use of antennas with a peak gain each of about 14.5 dB. The net gain (without adjacent beam interpolation circuitry) would be about 10-11 dB, and it was concluded that the concept was relatively complex and physically large for such a gain.

In general, for a lower gain requirement, the concept obviously has increased merit. Far fewer elements would be required since individual elements can have broad beams. In theory, a low gain system of about 4 to 5 dB can be provided with as few as four antennas of quadrant coverage characteristics. In practice, it takes a few more radiators to accommodate the coverage required for maritime satellite communications. Each antenna would be oriented to essentially provide coverage of a quarter of the hemisphere. The crossed-dipole over ground plane or cavity-backed spirals could be used as elements; however, the helical or conical spiral antenna offers pattern performance which can be more readily varied. The beamwidths of the crossed-dipole and cavity-backed spiral are very much fixed, whereas the helical antenna beamwidth can be adjusted by varying antenna length.

A five-turn, dual-wound helix would produce a symmetrical beam of 80° beamwidth at the half power points. The associated peak gain is slightly greater than 7 dB with respect to a circularly-polarized source, i.e., 7 dBic. The dual wound spiral is fed from the top end of the helix as shown in Figure 5-15. Either a Roberts or split balun could be employed to excite the dual-wound spiral. In each case, the two interwound spirals are excited in a balanced manner.

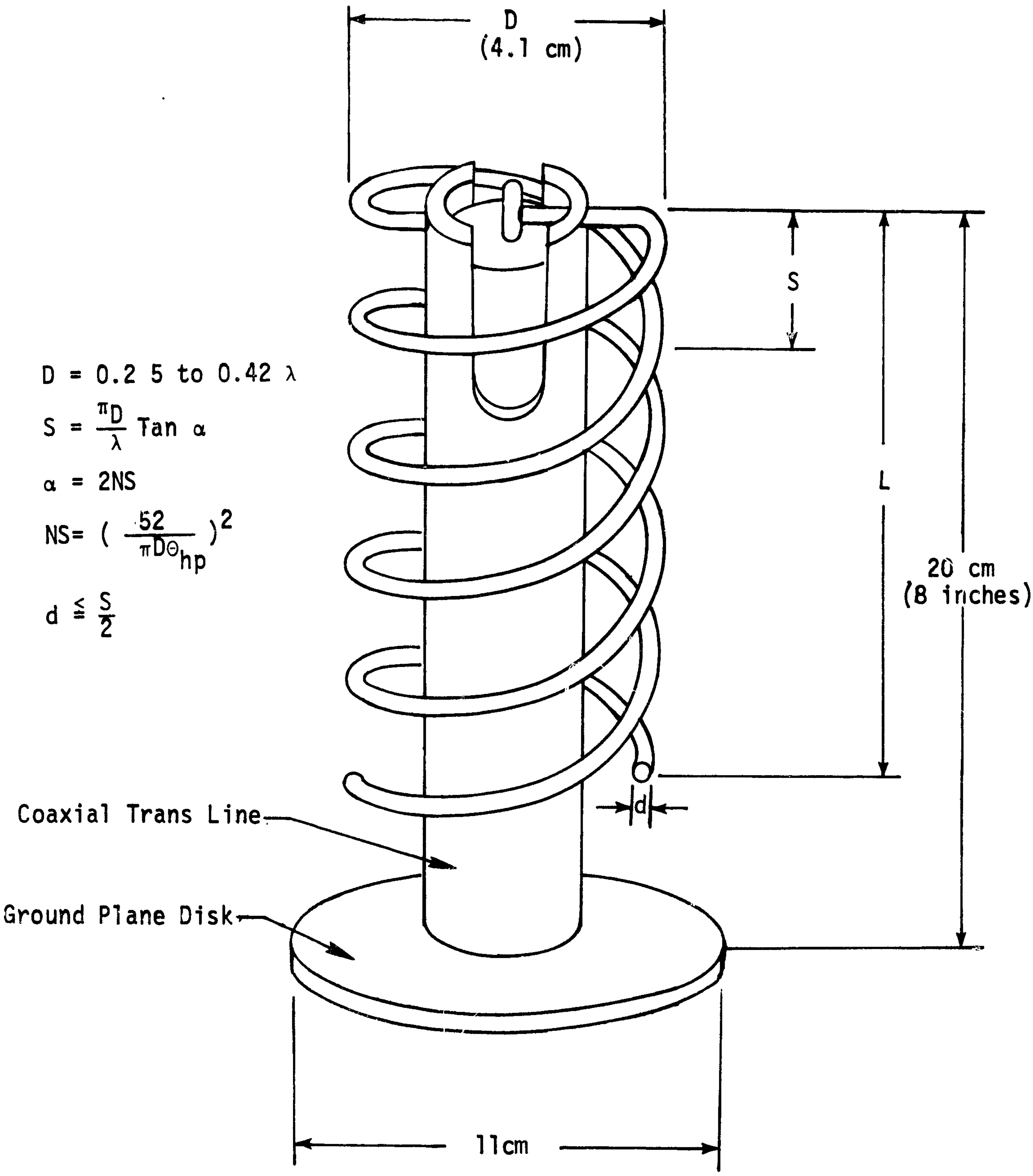

Figure 5-15 UNIDIRECTIONAL DUAL-WOUND HELIX ANTENNA

The feed line may be inserted through the center and may be used to support the helix by making it rigid.

Figure 5-16 shows how helical antennas may be employed to form a beam switching antenna for shipboard satcom application. The design employs eight helical antennas, and provides a minimum coverage gain of about 6 dBic. Similar to other designs, the most desirable location is about the mast, or on the port and starboard side of one of the upper bridge decks of the ship. By locating one antenna on each side of the ship, an unobstructed total coverage of greater than a hemisphere is obtained. Figure 5-17 illustrates the coverage possibilities. By separating the antenna elements into two groups, the tall structures such as smoke stacks or radar antennas would not obstruct the viewing angles of the antenna. Operation of the helical antenna beam switching system is schematically represented by Figure 5-18. Each helical element provides a broad 80° beam so that the motion of the ship can be easily tolerated. Remote switching would best be done by slaving to the ship's inertial references, i.e., the gyrocompass and a simple inclinometer, although automatic tracking is possible by comparing signals received in all beams. This example, with such wide beams, would not be amenable to conventional monopulse operation, although the analogous beam comparisons can be made. For example, a single receiver could be used to rapidly scan the outputs of all eight antenna ports to always indicate the best one. This would require a 3 dB power divider in the receive line prior to the communication channel switch (i.e., in Figures 5-1 or 5-2), with a separate switch for rapid scanning. As this 3 dB loss would be intolerable prior to receiver preamplification, the configuration of Figure 5-2 would have to be used, at least in the receive path. That is, eight diplexers and preselector/preamplifiers would be required. This merely serves to strengthen the

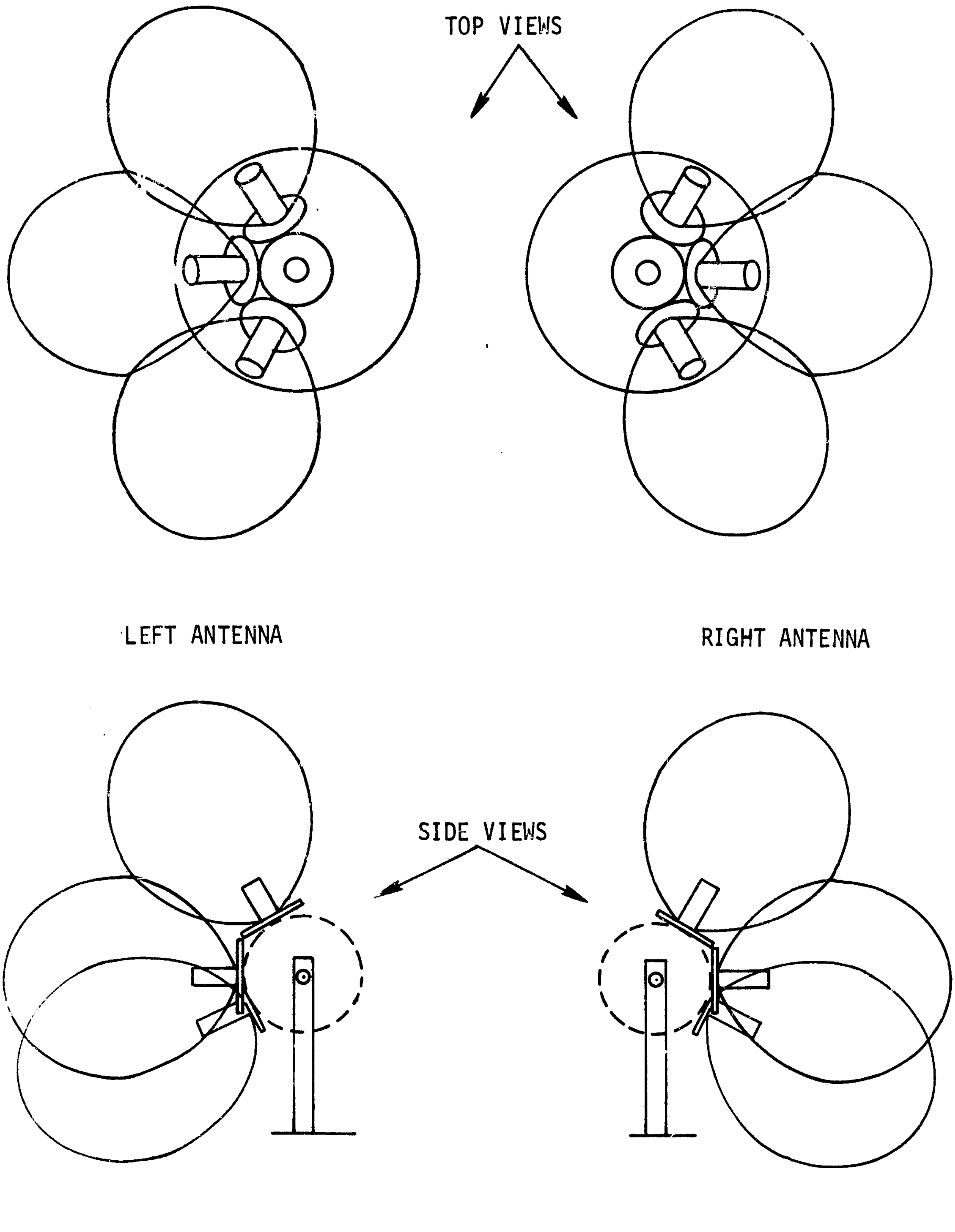

FIGURE 5-16 HELIX BEAM SWITCHING ANTENNA DESIGN

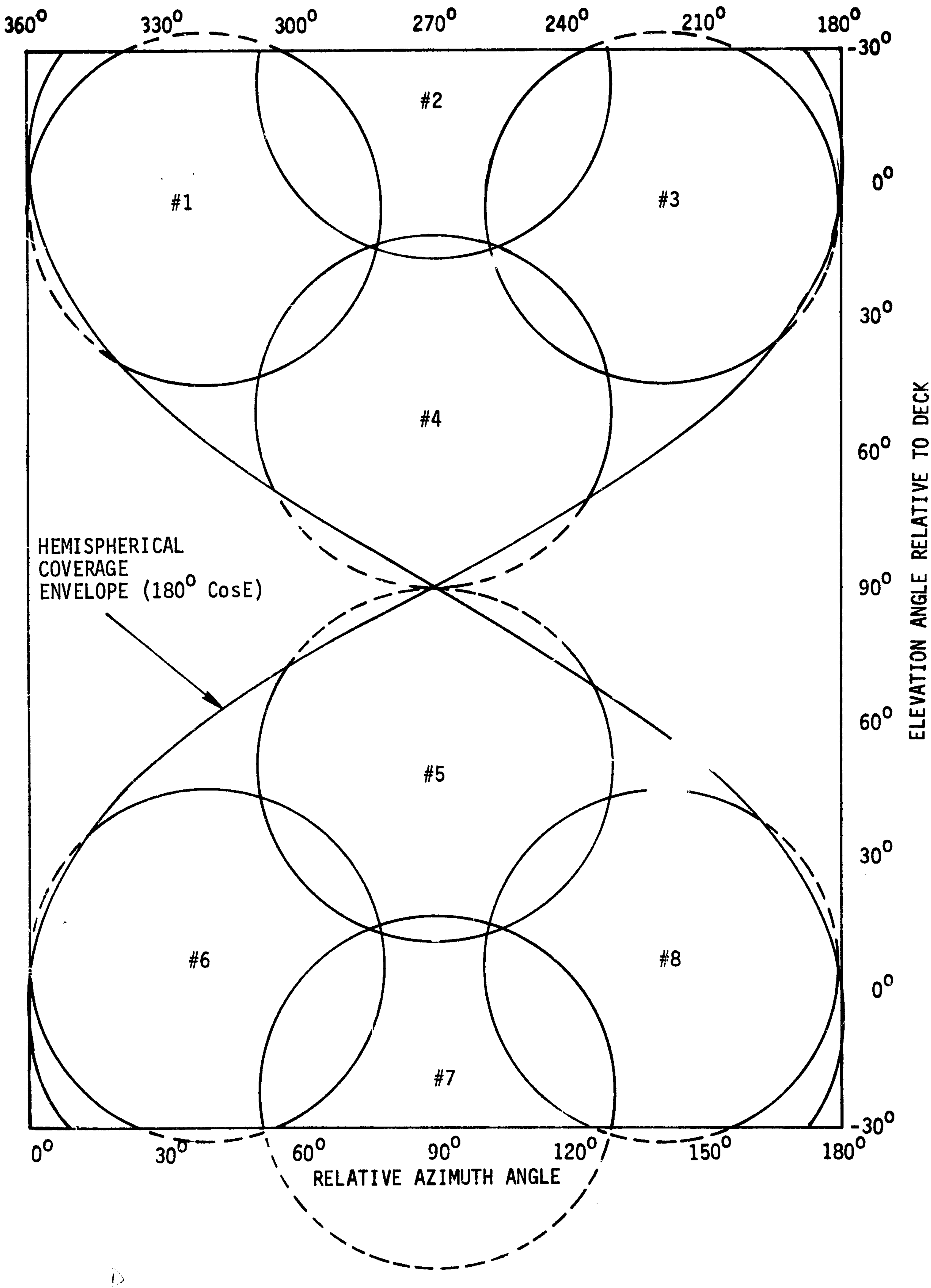

Figure 5-17 SWITCHED BEAM ANTENNA COVERAGE WITH 4-5 dB MINIMUM GAIN

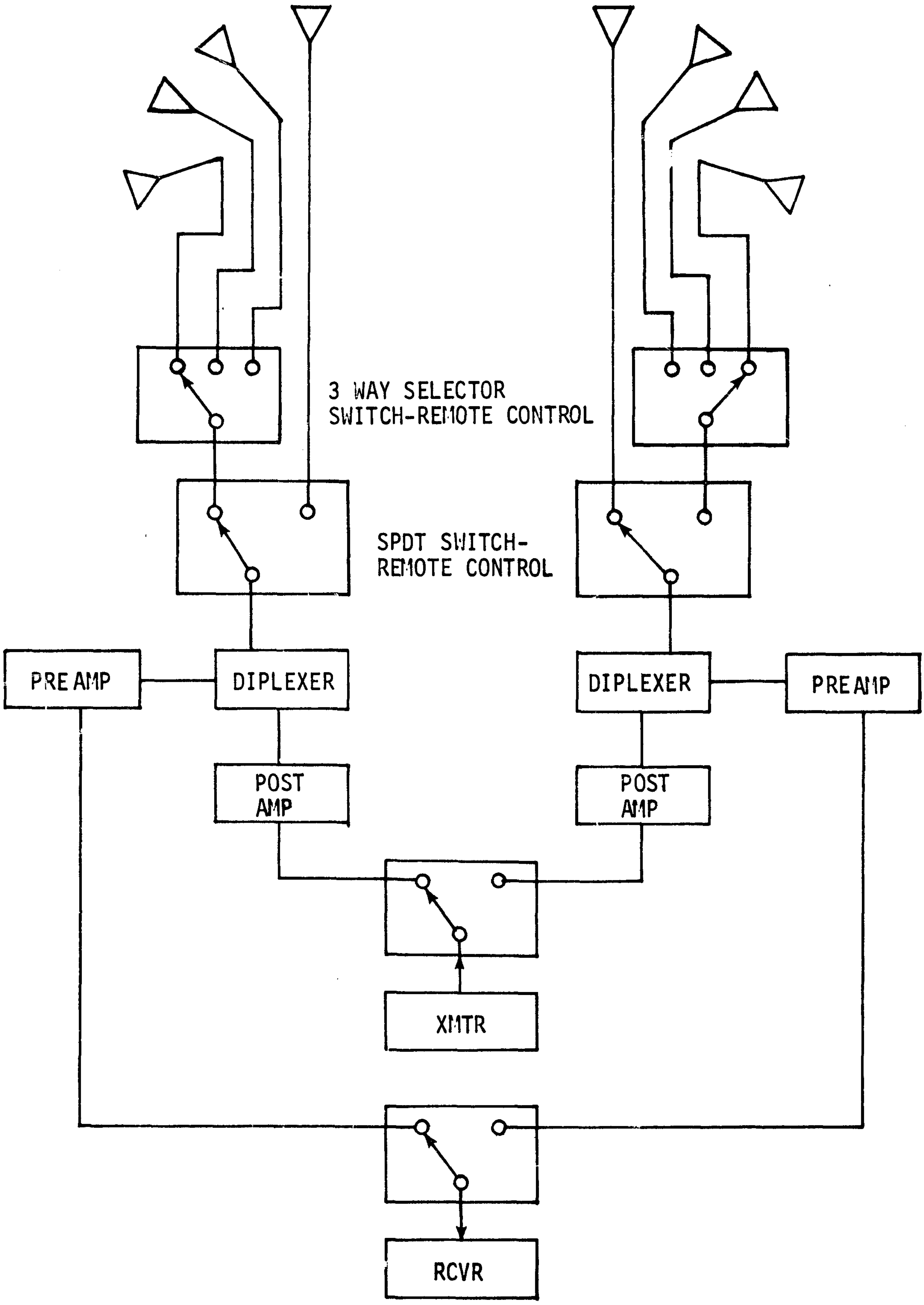

Figure 5-18 BEAM SWITCHING BLOCK DIAGRAM

conclusion that slaving is the more desirable means of antenna pointing for switched-beam antennas.

5.6 SUMMARY AND CONCLUSIONS FOR BEAM SWITCHING CONCEPTS

The preceding sections have served to indicate that the use of multiple antennas, or antennas with multiple beams, with beam switching, is feasible for the maritime shipboard application. In essence, the basic advantage of the switched-beam concept, that corresponding to no mechanical motion, is brought about by a large increase in RF circuitry. However, this, in itself, is not necessarily a critical limitation, especially when considering low-gain requirements.

It has been shown that a moderate gain requirement of 4-5 dBic can be easily satisfied by a very simple design consisting of eight helical antennas. A quadruple-eight element Butler array design of considerably more complexity can produce about 3-4 dB higher gain, depending on whether adjacent beam interpolation circuitry is included. Also, an example of two corner reflector arrays was shown to provide a gain coverage of 9 to 10 dBic for the ship. However, the latter design suffers major disadvantages in size and complexity.

Off-axis circular polarization performance can be best for the Butler array if curved turnstile or spiral elements are used. (See Section 4.11). The helical system and the corner reflector system both are limited to conventional transmitter/power amplifier arrangements. The Butler array permits the use of multiple low power transmitter power amplifiers, so that a certain amount of combining can be done by the array aperture itself.

For example, it was shown in Figure 2-5 that a 10 dB gain shipboard antenna permits use of a 100 watt shipboard transmitter power amplifier. An 8 dB (net) gain antenna would require a transmitter power of about 160 watts. These powers can be achieved by series and parallel combinations of transistors, i.e., a transistor array. In the quadruple

8-element linear Butler array antenna concept, each antenna array can be used to combine transmitter power outputs with one-eighth the power, e.g., 20 watts, feeding each array element. This would constitute a minor advantage of the switched beam array concept, although it is inherent that this advantage means use of multiple diplexers and receiver preamplifiers as well.

All systems are capable of mechanically sound design for the shipboard environment, by the use of plastic foam-filling and fiberglass radome covers. The effects of severe ice formation over the aperture is probably most serious for the Butler array and least deleterious to the helical system.

Table 5-2 summarizes the very general characteristics of the switched-beam example configurations considered. The most credible candidates are the first and last, i.e., the simple array of 8 helix antennas and the quadruple Butler 8-element array with beam interpolation.

As noted at the outset, the main disadvantage of the switched-beam category is complexity of RF circuitry. It should not be construed that this complexity reduces reliability, however, relative to a mechanically-pointed beam. In theory, the array approach may be most reliable. However, the increased circuit complexity does mean increased shipboard terminal cost.

A good deal of the additional RF circuitry, while functionally complex in block diagrams, can readily be done in stripline at 1600 MHz. This permits the required functions to be implemented fairly cost-effectively when manufactured in large numbers. No precise knowledge of costs is known, but it is estimated that the cost of a switched-beam antenna subsystem would be considerably larger than a mechanically pointed, single-beam antenna of equivalent gain. This conclusion does change with the gain requirement; at very low values of gain such as 3-5 dB net, the

TABLE 5-2 SUMMARY COMPARISON OF EXAMPLE SWITCHED-BEAM ANTENNA CONCEPTS

Switched-Beam Example / Performance Parameter	Simple 8-Helix Combination	Array of 12 Corner Reflectors	34 Separate Short-Backfire Antennas	Dual Butler Linear 8-Element w/ 4 Helices	Quadruple Butler (Linear) 8-Element Array w/ Beam Interpolation
Minimum Gain, dBic	4-5	6-7	10-11	6-7	9-10
Peak Gain, dBic	7-8	9-10	14.5	10	10-11
Basic Beam Shape	Circular, 80°	Fan, 30°x120°	Circular, 33°	Fan/Circular, 20°x130°	Fan, 20°x130°
Physical Size	Medium	Very Large	Very Large	Large	Large
Relative Axial Ratio for C.P.	Good	Poor	Good	Fair	Best
Potential Environmental Degradation	Small	Moderate	Small	Icing Problem	Icing Problem
Complexity, Relative Cost	Lowest	Medium	High	Higher	Highest

reverse would probably be true.

It was estimated in Section 3.7 that the cost of a mechanical pointing subsystem (not including automatic tracking or slaving peculiar equipment) would be about $2800 in production. The cost of a simple low to medium gain (3-15 dB) antenna radiator, when weather protected, is estimated to be about $150. Thus, the total cost of a mechanical system should be less than $3000. While this would indicate that as many as 20 antennas without mechanical pointing could be used, it is estimated that this number would be reduced to about 10 to 12 when the beam-switches and related logic are included. When additional installation costs are included for the significantly increased physical size of the antenna system, it is believed that the trade-off number of radiators is closer to 6 or 8. This corresponds to antenna net gains on the order of 4-5 dB.

In summary, it can be concluded that the switched beam concept is especially suitable to low-gain antenna system requirements ($\leq$6 dB), and that large separations between individual radiators are not feasible unless multiple RF electronics groups are employed, and that beam-switch control is best accomplished by slaving to appropriate shipboard references.

SECTION 6

ELECTRONICALLY-STEERED BEAM ANTENNAS

6.1 THE ELECTRONIC SCANNING CONCEPT

In contrast to switching between fixed beam positions, the linear and planar arrays discussed in Sections 4 and 5 can be continuously steered in direction by the continuous control of phase-shifters. Consistent with common terminology, the electronically-steered beam arrays are referred to as phased arrays. Over the last decade, phased array technology has received a great deal of attention.(63) The primary application for the technology has been radar, for search and tracking.(64)

The phased array is a versatile antenna which can be designed to perform a wide variety of functions in communications as well as the radar field. The antenna can perform very rapid scanning without physically moving the antenna aperture. This is done by rapid changing of the phasing of each antenna element to provide a collimated beam in the desired direction. Because the antenna elements are individually controlled, various beam shapes, including the formation of simultaneous beams, are possible. The major disadvantages of phased arrays are high cost and complexity resulting from the many additional components relative to conventional antennas. They are most effective when rapid movement of a single beam is required.

The most common geometrical forms of array antennas are the linear and

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(63) Hansen, R.C. "Microwave Scanning Antennas," 3 Volumes, Academic Press, New York, 1966.

(64) Hardeman, L.J. "Phased Arrays Scan Rapidly Towards Growth in the 70's" Microwaves, June, 1970.

planar arrays. The linear array design provides one-dimensional steering of fan shaped beams, while planar phased arrays can be designed to provide two-dimensional steering of pencil beams.[(65)] Other types of array configurations are possible. For example, the elements can, in theory, be arranged on the surface of a cylinder to generate 360° of coverage. The same coverage may be also obtained with a number of linear or planar arrays. An array may be also designed onto the surface of a sphere, or an object of almost any shape, but each element must be carefully phased to produce plane waves.

For the subject shipboard application, fan beam steering is most desirable since the coverage can be satisfied by one dimensional steering. For a moderate value of required gain, such as 10 dB, a typical example corresponds to the use of six linear arrays, with three on each side of the ship. Such an arrangement is illustrated in Figure 6-1. As in the switched beam category, the array should be tilted about 30° in elevation for best coverage, including ships roll, by the broad element pattern. Each array contains 8 elements and provides 60° steering. The individual linear array arrangement can be as discussed in Sections 4.5 and 5.4. The elevation angle coverage is fixed, at about 120°-130° for typical large ships. The azimuth angle beamwidth is quite narrow, i.e., about 10-15 degrees. A number of element designs can be employed in the array to yield the required beamwidth. Since the crossed-dipole is the simplest to produce, its use will be assumed in the following discussions.

It may be noted at the outset that the configuration shown in Figure 6-1 is in fact a combination of electronic steering and beam switching,

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(65) Hering, K.H. "Optimization of Tilt Angle and Element Arrangement for Planar Arrays" Microwave Journal, January, 1971.

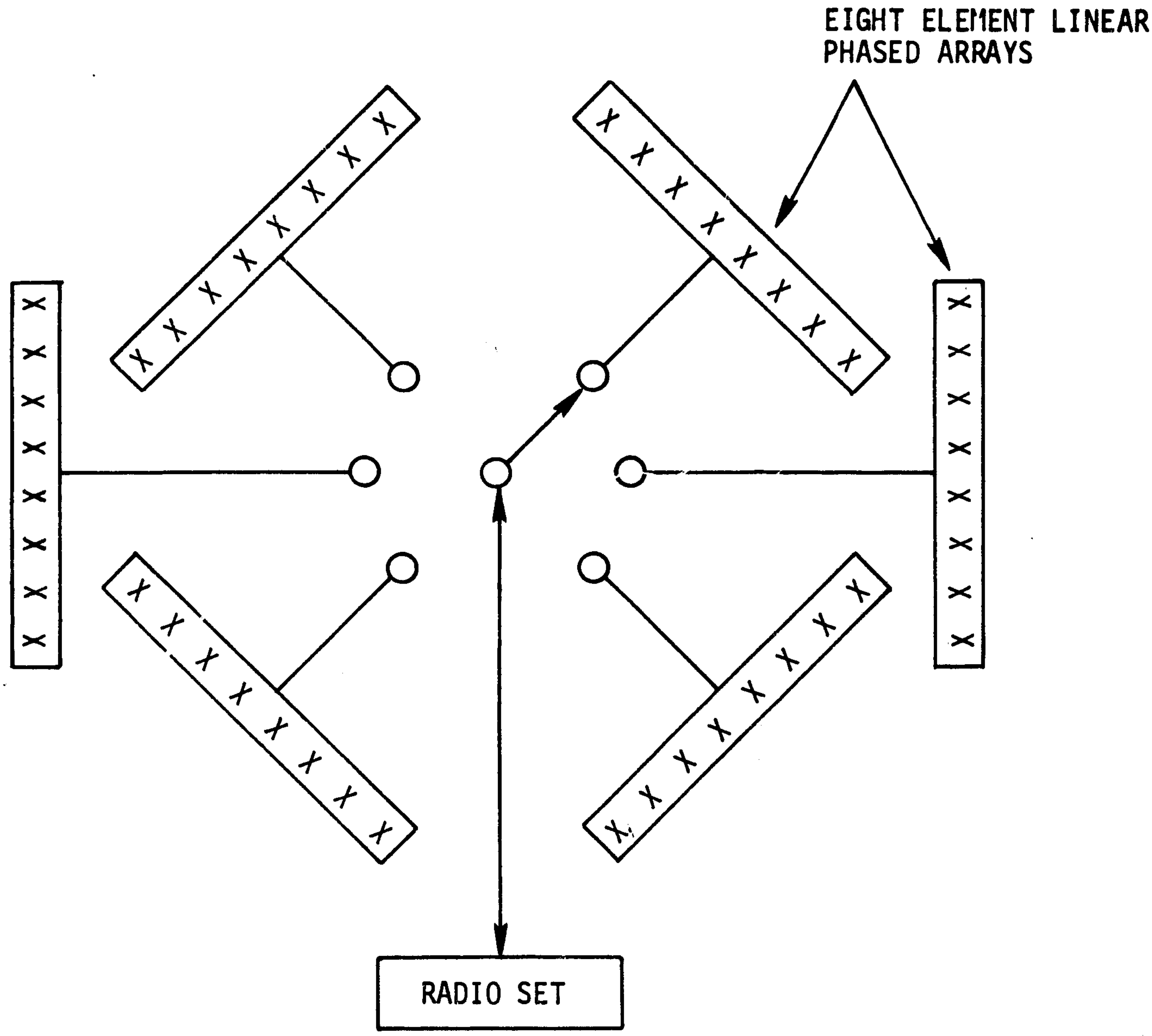

FIGURE 6-1 ELECTRONICALLY-STEERED ANTENNA CONFIGURATION FOR 10-13 dB GAIN, WITH SWITCHING BETWEEN INDIVIDUAL PHASED ARRAYS

where the switching is between "groups" of beams. It is almost axiomatic in this category that some switching between groups is required, in that it would be extremely difficult to provide the required greater than hemispherical coverage with a single phased array. The gain and circular polarization of an array in the end-fire direction (parallel to array aperture) are typically severely degraded. Also, the same considerations illustrated in Figure 5-1 and 5-2 are pertinent in this category; that is, a variety of alternatives in phase-shifter and group switch locations exist, with varying degrees of RF circuit complexity and performance.

The linear array which generates a single beam can be converted to a multiple beam antenna by attaching additional phase shifters to the output of each element. This permits a number of beams to be generated simultaneously from the same aperture. The multibeam feature could be used for tracking purposes to steer the communication beam to the direction of the strongest signal. Low-noise amplifiers may be placed between the antenna elements and the phase shifting networks to amplify the incoming signals and compensate for losses in the phase-shifting network. By definition, diplexers would also be used then between elements and phase-shifters, so that twice as many phase-shifters, i.e., 96, would be required. Steering can be also accomplished via signals from the gyrocompass and an inclinometer. In the latter approach, the array design is greatly simplified since the individual element requires one phase shifting device only. It is to be noted that a vertical reference would be required in this narrow azimuth beam example, to compensate for azimuth errors caused by ship roll, as described in Sections 2.4 and 2.5.

6.2 LIMITATIONS IN ARRAY SCAN ANGLE

As noted, scanned arrays become more difficult to design as the scan angle away from broadside increases beyond ±45. This is due to two problems - 1) maintenance of good circular polarization and 2) maintenance of good impedance match. Due to mutual coupling between array elements, the input impedance can develop high mismatch values at specific off-axis beam scan angles. These "blind spots" can cause more than 50% reflection of the transmitter power. A simple model of the reflected power is [66]

$$R(\Theta) = [\tan(\Theta/2)]^4 \qquad (6\text{-}1)$$

where $R(\Theta)$ is the reflection coefficient and Θ is the scan angle from broadside. This equation is based on a theoretical "current sheet" array which has a matched impedance at broadside and has only a resistive mismatch at other scan angles. Knittel[67] states that real arrays not utilizing the complexity of scan dependent reactive matching networks will have aperture reflections two and three times $R(\Theta)$. Thus, the above $R(\Theta)$ expression is considerably optimistic. Accordingly, Knittel presented the following table summarizing the effect of scan angle design on gain, including the aperture projection scan loss. The element spacings given are optimized for maximum gain without grating lobes.

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(66) Wheeler, H.A. "Simple Relations Derived From A Phased Array Made of an Infinite Current Sheet" 1964 IEEE Antennas and Propagation International Symposium Digest, pp. 157-160.

(67) Knittel, G.H. "Choosing the Number of Faces of A Phased-Array Antenna for Hemisphere Scan Coverage" IEEE Trans. on Antennas and Propagation, November 1965.

TABLE 6-1

EFFECT OF SCAN ANGLE ON GAIN IN A PHASED ARRAY

Max. Scan Angle	63°	55°	47°	41°
Element Spacing	0.628λ	0.691λ	0.679λ	0.700λ
Max. Power Reflected	14%	7%	4%	2%
Broadside Beamwidth / Max. Scan Beamwidth	0.45	0.57	0.68	0.76
Broadwide Gain* / Max. Scan Gain	4.1 dB	2.8 dB	1.9 dB	1.3 dB

It is obvious then, that smaller scan angles produce better performing arrays. However, smaller scan angles requires more arrays to cover a hemisphere. Thus the usual trade-off exists of more (costly) equipment for better performance.

The use of four linear fan beam arrays for 10 dB minimum gain over the hemisphere requires a maximum scan angle of 45°, i.e., close to the 41° value in Table 6-1. This number of arrays then would maintain the scanned-beam gain performance to within 1 to 2 dB of that realized at broadside. The six array configuration depicted in Figure 6-1 would obviously reduce the degradation, since the maximum scan angle would be decreased to ±30°. As the gain improvement may be small, the use of four linear 8-element arrays is probably a good compromise. Extrapolating, the use of just three arrays would require scan angles of ±60°, with a scan loss of 4.1 dB. Thus, the fourth array is clearly warranted.

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* Realized Gain in the presence of mismatch loss.

6.3 PHASE SHIFTER DEVICES

A critical element in phased arrays is the phase-shifter itself. Two of the most widely used electronically-controlled, phase shifting devices in array application are the ferrite and diode phase shifters. These devices can be designed to have phase variations in a continuous fashion or in discrete states. The ferrite devices are known for low loss performance and low driving power requirements in waveguide design. Their use is firmly established for frequencies above 3000 MHz. On the other hand, the diode phase shifters appear to be generally superior for frequencies below 3000 MHz in size, weight, power handling and temperature considerations.[68] The following paragraphs describe the characteristics of diode phase shifters.

6.3.1 Diode Phase Shifter Control Elements

The control elements for diode phase shifters are varactors and pin diodes. These elements can be integrated with transmission lines, circulators, and hybrid couplers to form a variety of phase shifters with continuous or stepped phasing. The continuous phase shifter would make use of a continuous variable capacitance from the varactor diodes to alter the phase of microwave energy passing through them. On the other hand, the step type of phase shifters would utilize essentially the short and open circuited impedances of the pin diodes to produce the required discrete phase changes.

Of the two basic types of control elements, the varactor is limited in power handling capability because capacitance changes produced by a bias change are nonlinear. Therefore, the RF voltage swing in the varactor must be restricted so that the average capacitance observed by the RF

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(68) Mortenson, K.E., "Microwave Semiconductor Control Devices", Microwave Journal, May 1964

signal is approximately the same at all bias points. A tolerable RF voltage swing is approximately 10 percent of the breakdown voltage.

6.3.2 Diode Phase Shifter Power Handling Capabilities

The power handling capabilities of varactor phase shifters may range from 100 watts average at UHF to a few watts at X-band with special varactor diodes, while the pin diode unit can be fabricated to handle much greater power at these frequencies with a smaller insertion loss. The latter can handle five to ten times the power of the former, and should thus be used in the subject L-band application. In peak power comparison, the pin diode construction exceeds the varactor design by several orders of magnitude. A comparison of two types of phase shifters by Mortenson[68] is given in Table 6-2. This comparison clearly indicates that pin diodes are superior to varactors in the areas of higher power handling, smaller insertion loss, and greater phase range.

6.3.3 Pin Diode Phase Shifter Designs

Several basic designs using pin diodes to form phase bits are shown in Figure 6-2. In each case, the diodes are incorporated in such a manner that a change in diode impedance causes a change in RF signal transmission. These diodes are operated in two alternative dc biases to provide near open and short circuited conditions. An increment of phase shift is achieved by reverse and forward biasing of the diodes. The configurations are described below, respectively.

Reflection Time Delay Configuration

Figure 6-2a and 6-2b are phase shifters operated on reflection principles. The diodes are used to alter the short circuited positions at

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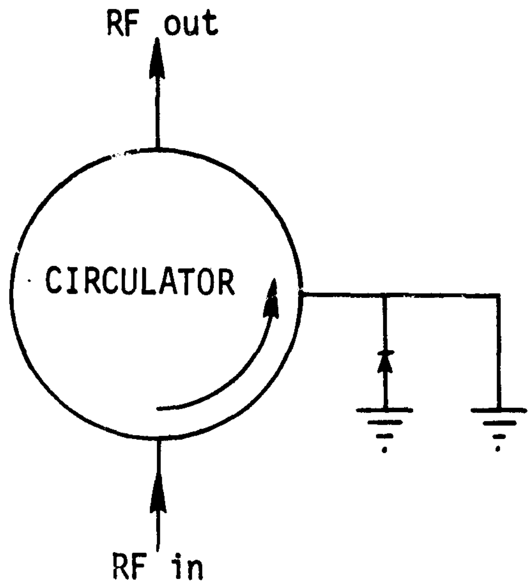
[68] Mortenson, K.E., Op. Cit.

TABLE 6-2

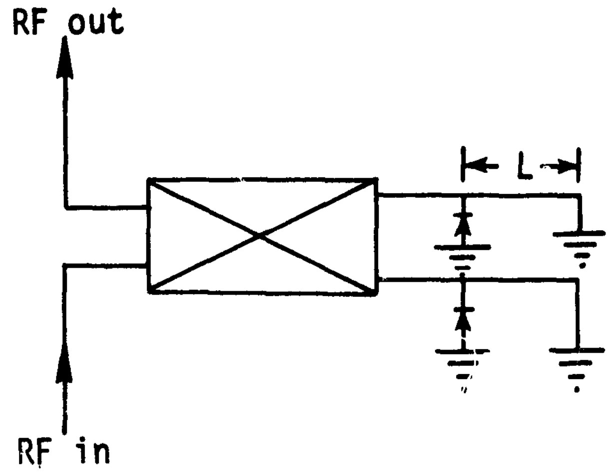
COMPARISON OF STEP PHASE SHIFTERS AND CONTINUOUSLY VARIABLE TYPE PHASE SHIFTERS*

Characteristics	Step-Type Phase Shifter	Continuously Variable Type Phase Shifter
Phase	Incremental, Step values unlimited-dependent only on line lengths and insertion loss limits (as much as 360° steps)	Continuous, range limited by capacitance change and insertion loss limits (up to 220°/unit)
Insertion Loss	Least dependent upon diode - 0.4 dB at L-band	Dependent upon diode - 1.5 dB at L-band
Peak Power	Less RF voltage or current swing restrictions - 50 kW	RF voltage swing restricted by capacitance change - 100 W or less
Average Power	100 W at L-band	25 W at L-band
Phase Power Level Sensitivity	None (within power rating)	Dependent upon RF voltage swing compared to breakdown of varactors
Phase Stability:		
a. Temperature	Independent up to about 150° C	Dependent upon capacitance change - 0.1°/°C
b. Control	Not critical as long as forward or reverse bias is maintained, no hysteresis effects	No hysteresis effects
Speed	Moderate, less than 100 ns	Highest, nanosecond range

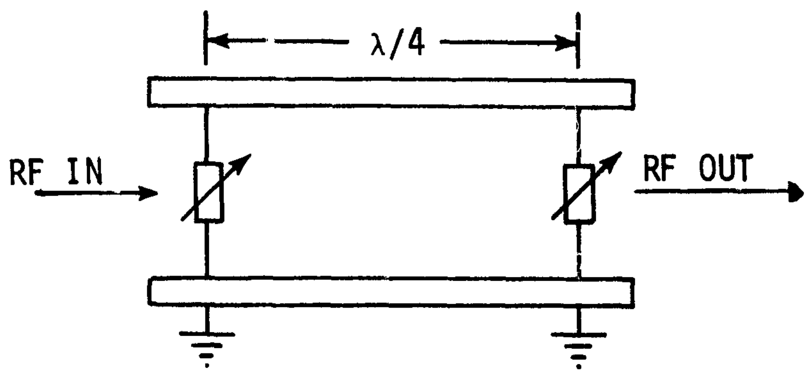
* Diodes Operating in the Reflective Mode on 3-dB Hybrid Coupler



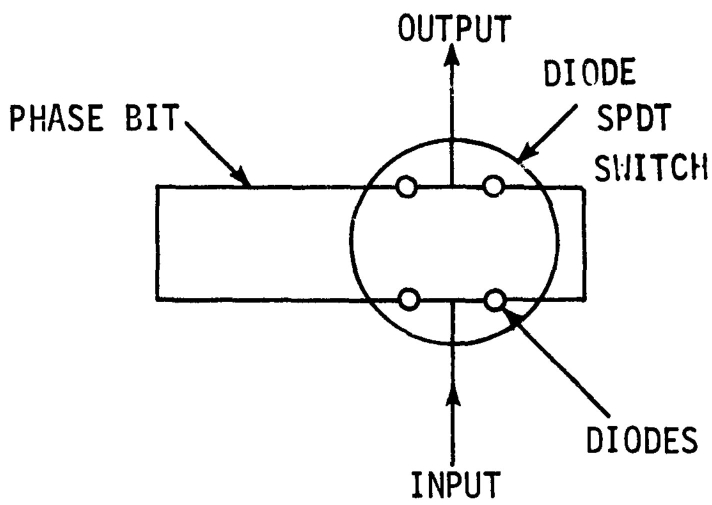
A. Transmission-Reflection Time Delay Phase Shifter



B. Bilateral Transmission Reflection Time Delay Phase Shifter



C. Loaded Line Phase Shifter



D. Switched Line Length Phase Shifter

Figure 6-2 ALTERNATIVE DIODE SHIFTER DESIGNS the end of the lines to the diode position by applying approximately 100 ma of dc current to each diode. The phase change produced by this design is approximately twice the line length between the diode and its short circuited termination. The design indicated in Figure 6-2b is capable of bilateral performance but symmetrical terminations must be maintained on both coupling arms. The line lengths must be equal. Also, the diodes must have equal characteristics and be positioned at the same distance away from their terminations. This is to ensure that the waves reflected from the coupling arms are self cancelling at the input terminal and that all the reflected waves are summed up at the output port to produce a low VSWR.

Loaded Line Configuration

The design of Figure 6-2c, referred to as a loaded-line phase shifter, uses quarter-waved-spaced diode pairs to perturb the phase of the transmission coefficient. The diode pair can take on either of the on and off states and the total transmission phase shift is determined by the number of pairs of diodes in the line and their states. The total phase shift for this type of phase shifter is approximately equal to the phase shift of a section, times the number of sections in the line. The largest measured phase shift per diode pair reported is 23°; hence, many diode pairs are required to obtain a phase shifting range of 360°.

Switched-Line Length Configuration

Figure 6-2d is an illustration of a switched-line length design where diodes are used to switch incremental line length. This design consists of two SPDT switches and two line-lengths for each bit. The number of diodes required per bit is four. In an ideal construction, this design can handle twice the peak power of the hybrid-coupled-bit design in Figure 6-2b, since no reflected waves occur in the switched-line circuit and the doubling of the RF voltage is precluded. The loss is approximately identical to a 180° phase bit of an optimized hybrid-coupled unit. Therefore, the total loss of a four-bit switched-line phase shifter is four times the loss of the 180° hybrid-coupled bit and 2.28 times the same loss for the four bit hybrid-coupled design.

6.4 DIGITAL VS ANALOG (CONTINUOUS) PHASE SHIFT SCANNING

6.4.1 Digital Phase-Shifting

Two types of phase-shifting in arrays are possible; digital and analog (continuous). In the digital approach, the phase shift and thus the antenna beam motion is in discrete steps. The accuracy with which a phased array beam can be pointed thusly is determined by the minimum phase step size.

The number of phase bits which are required to provide a certain beam-position accuracy in a phased array can be ascertained as follows.(69) An approximate formula which relates phase bits "n" to array elements "N" and fraction of the $2/\pi$ (voltage) beamwidth Θ_V is given by

$$n \approx \log_2 N/\Theta_V \tag{6-2}$$

For example, if an 8-element array is assumed and the desired beam-position is $\Theta_V = 1/2$ (beam-position is controlled to half of the $2/\pi$ voltage beamwidth), then according to the above expression, the number of binary phase bits required is

$$n \approx \log_2 8/\tfrac{1}{2} = 4 \tag{6-3}$$

The number of beam positions available is $2^n = 16$ beams. Half of these beam positions are in the first quadrant, and half are in the second. Some of the positions will form grating lobes. Of course, the directivity and beamwidth are still determined by the size of the phased array. If the beam-position is controlled to $2/\pi$ voltage beamwidth, the required number of binary phase bits is three (for

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(69) MIT-Lincoln Laboratories Technical Report No. 236, "Phased Array Radar Studies." November 13, 1961.

example, 45°, 90° and 180°).

In a situation where the actual number of phase bits available is less than the number of phase bits required, an error in beam position results. An approximate formula which expresses the worst beam pointing error ($\Delta\Theta$) in terms of the $2/\pi$ beamwidths of a uniformly illuminated large array is

$$\frac{\Delta\Theta}{\Theta_V} \approx \frac{\alpha\, 2^{2b}}{2\pi N} \times 100\%$$

where α = required phase quantization
b = $n - m$
n = no. of required phase bits
m = no. of available phase bits
N = no. of array elements

In array practice, it is not necessary to quantize phase to an accuracy of $2\pi\,(\Theta_V/N)$ radians steps to obtain N/Θ_V beam positions. In general, n should be greater than m. In an example given in the referenced study[69] for a uniformly driven 64-element array, the required number of phase bits was 7 but the available phase bits was 4. The pointing angle error was determined to be less than 2%.

6.4.2 Continuous (Analog) Phase-Shift Scanning

Unlike the digitally-stepped phased array, the beam position of the analog array can be continuously steered by changing the bias voltage across the varactor diodes of the phase shifter to obtain phase shift. Except for

(69) MIT-LL Technical Report #236. Op. Cit.

phase shifter implementation, the analog array physical characteristics are identical to those of the digital array. The continuous phase-shift scanning array efficiency is slightly less than that of the digital system due to higher loss in the phase shifter. Thus, it may generally be concluded that the digital stepping approach is a superior selection for the L-band shipboard application.

6.5 CONCLUSIONS ON ELECTRONICALLY-STEERED ANTENNAS

Given the phased array concept, one of the primary considerations in the design, complexity and performance of the antenna is the location of the phase shifters in relation to the array elements. The various locations possible, with the corresponding array circuitry requirements, are indicated in Figures 6-3, 6-4, 6-5 and 6-6. Other configurations are possible; however, the four shown illustrate the basic variations. Obviously, these figures show only one eight-element array, while four are required for complete coverage. The utmost in circuit simplicity would exist if the switch used to select one of the four arrays were inserted between the combiner/divider and diplexer in Figure 6-6.

The arrangement shown in Figure 6-3 utilizes separate apertures to perform the transmit and receive functions. The configurations of Figures 6-4, 6-5 and 6-6 use common apertures, but with varying amounts of RF circuitry. Table 6-3 summarizes a comparison of the four basic configurations.

The number of components noted in the table refer to the complete antenna system consisting of four 8-element phased arrays with adjacent electronically steerable fan beams. A fifth category is included in Table 6-3 to distinguish between the use of four or one diplexers, according to the location of the switch used to select arrays.

In terms of power efficiency, it is clear that the first configuration, Figure 6-3, is superior. The antenna gain provided by the array aperture would be realized in operation. However, a total of 8 array apertures are required, with 64 elements, 64 phase shifters and 32 each of low-noise receiver preamplifiers and transmitter power amplifiers. The second configuration, Figure 6-4, is just slightly less efficient,

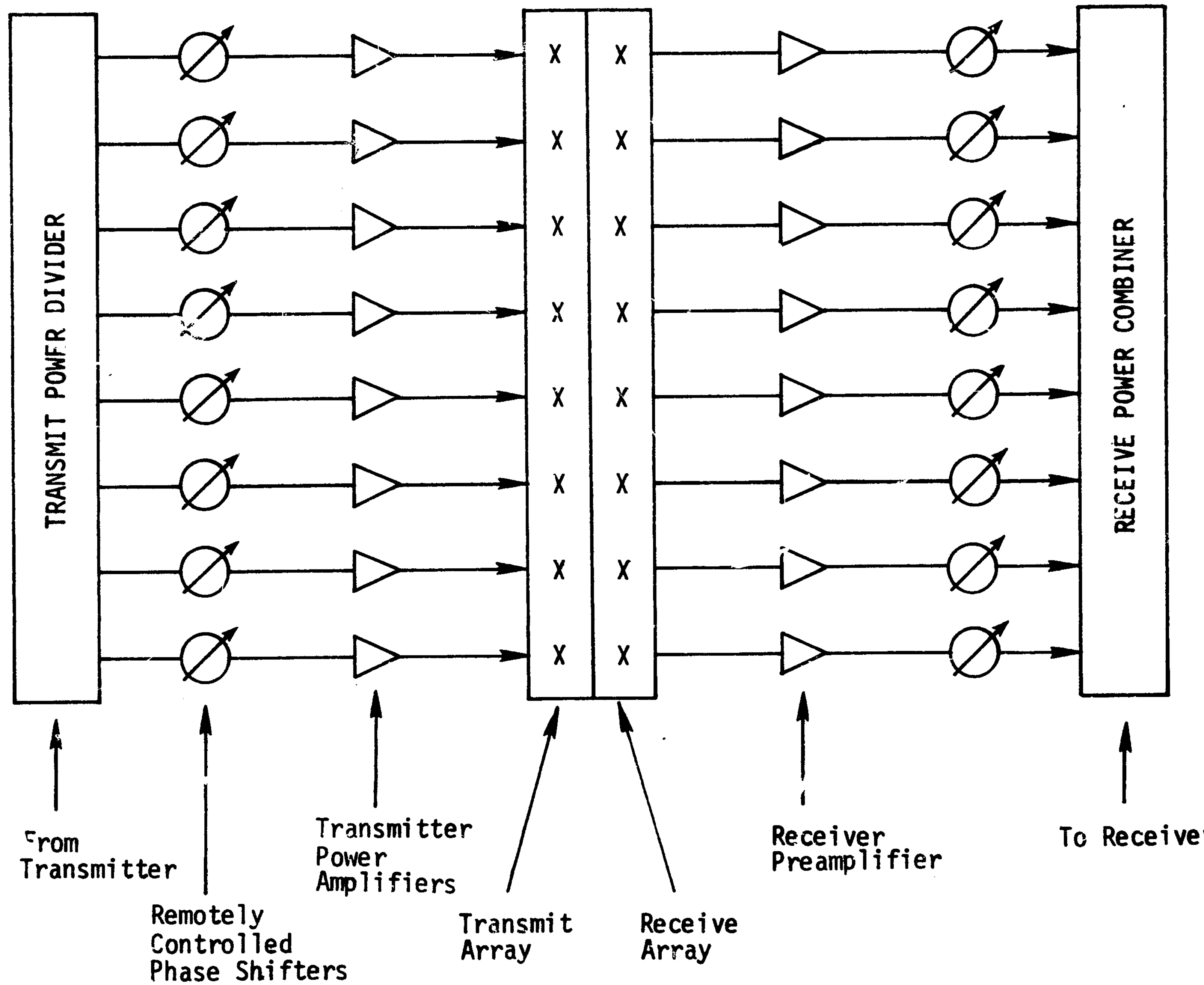


Figure 6-3 8-ELEMENT PHASED ARRAY WITH SEPARATE TRANSMIT AND RECEIVE APERTURES

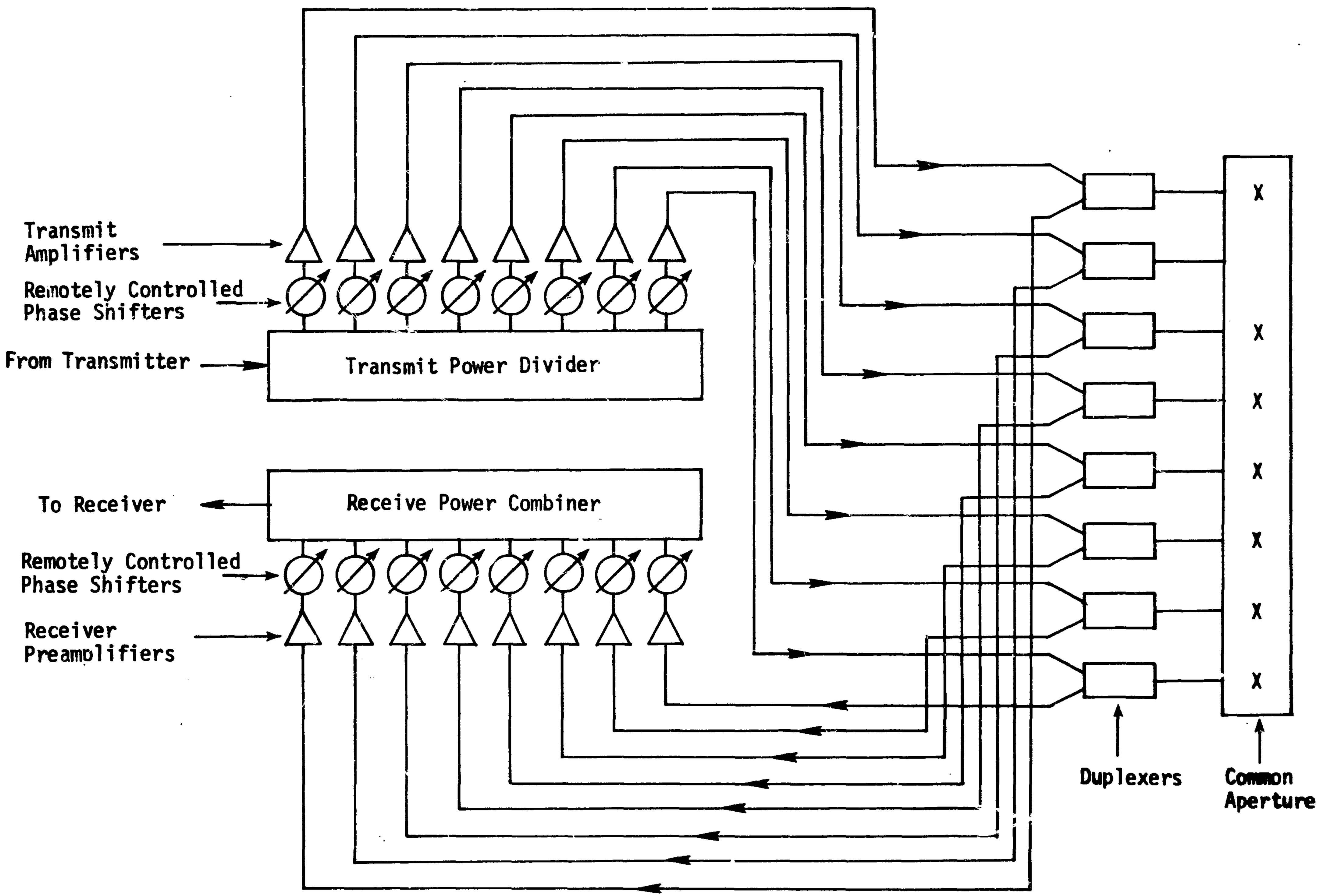


Figure 6-4 8-ELEMENT PHASED ARRAY WITH SHARED APERTURE BUT INDEPENDENT TRANSMIT AND RECEIVE BEAMS

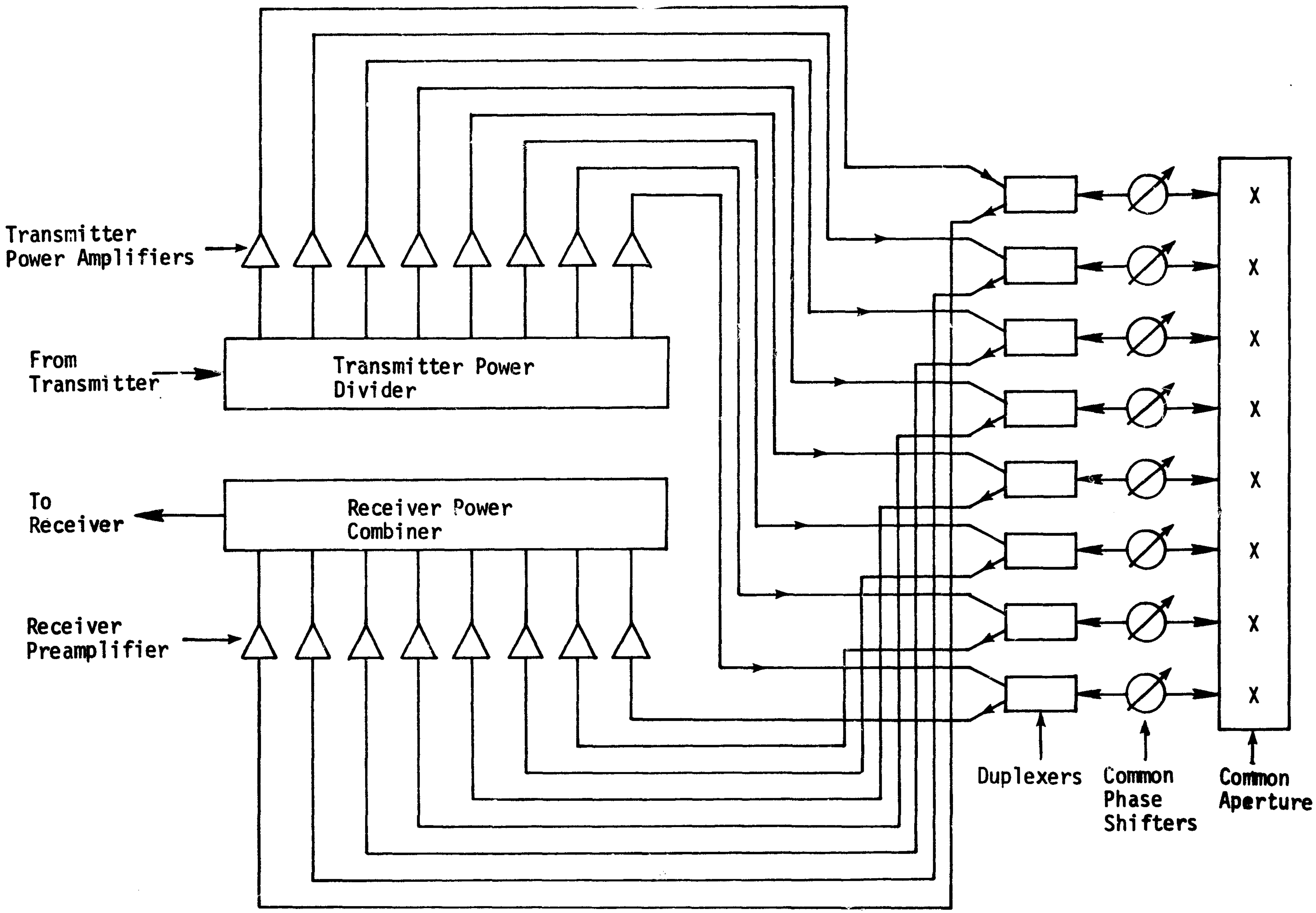


Figure 6-5 8-ELEMENT PHASED ARRAY WITH COMMON APERTURE AND COMMON BEAM STEERING

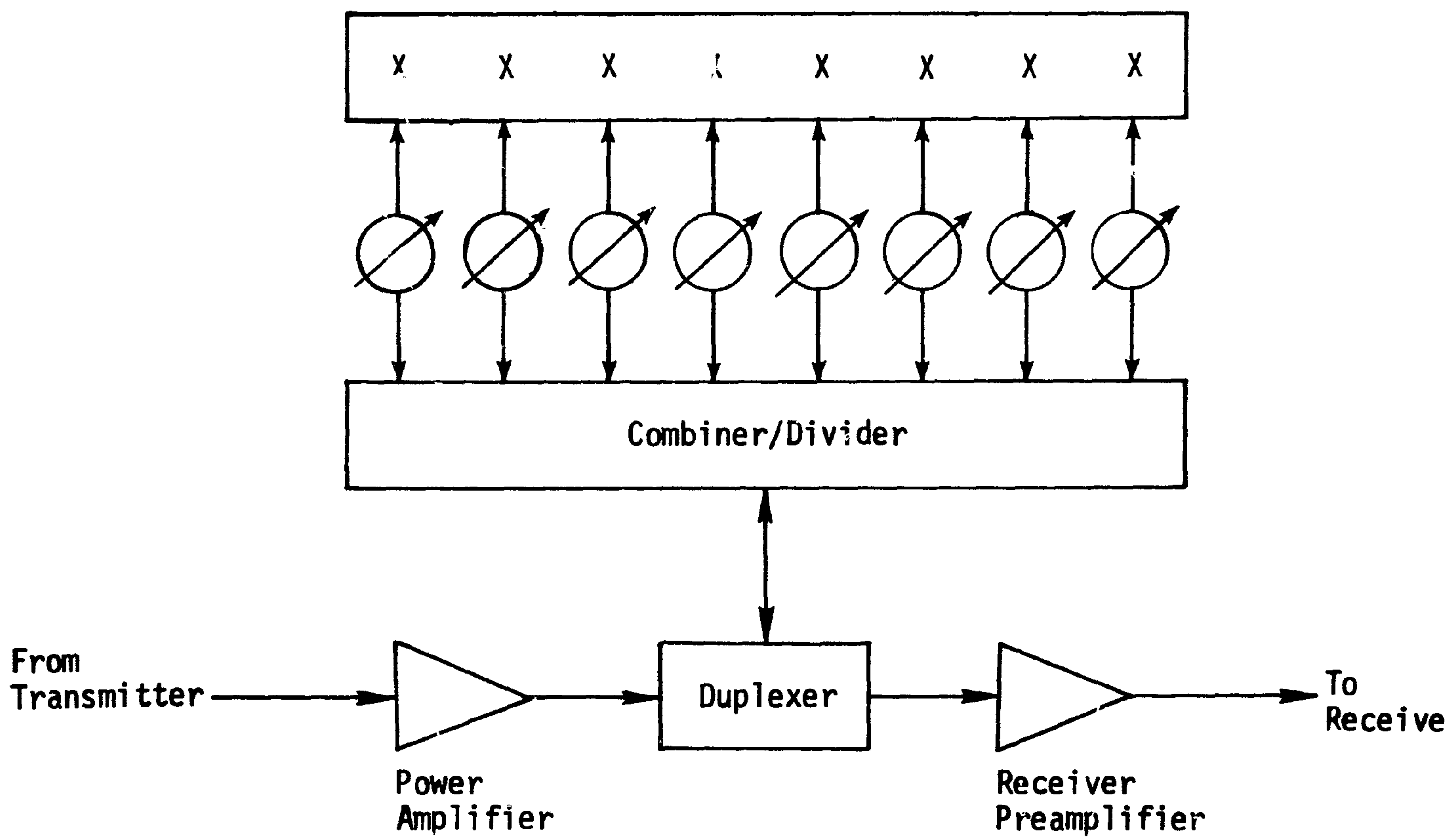


Figure 6-6 8-ELEMENT PHASED ARRAY WITH COMMON APERTURE AND SINGLE TRANSMIT/RECEIVE DUPLEXER

TABLE 6-3 SUMMARY COMPARISON OF PHASED ARRAY CONFIGURATIONS

CONFIGURATION	NO. OF APERTURES	NO. OF PRE & POST AMPLIFIERS	NO. OF PHASE SHIFTERS	NO. OF DIPLEXERS	EFFICIENCY	TRANSMIT/ RECEIVE BEAM	VOLUME	WEIGHT
Separate Transmit & Receive Apertures (Figure 6-3)	8	32 of each	64	None	Good	Independent	High	High
Common Aperture, 16 Phase Shifters per Array (Figure 6-4)	4	32 of each	64	32	Fair	Independent	Medium High	Moderate
Common Aperture, 8 Phase Shifters, 8 Diplexers per Array (Figure 6-5)	4	32 of each	32	32	Poor	Common	Medium High	Moderate
Common Aperture, Single Diplexer per Array (Figure 6-6)	4	4 of each	32	4	Very Poor	Common	Medium	Lower
Common Aperture, Single Diplexer with 4 Arrays	4	1 each	32	1	Poorest	Common	Low	Lower

e.g., -0.5 dB, but requires one-half the number of apertures and elements in a trade for 32 diplexers. The third configuration, Figure 6-5, is considerably less efficient due to phase shifter insertion losses. Diode phase shifters are quite lossy, typically with 2 to 3 dB insertion losses. Thus the third configuration is clearly undesirable since low efficiency results without a major decrease in the number of RF components.

The fourth configuration, Figure 6-6, is the least complex, especially if the 4-way switch between arrays is inserted between the phase shifters and a single diplexer for all apertures. In addition to having very low power (gain) efficiency, the latter configuration does not provide for "graceful" transmitter failure. In the first three arrangements, the transmitter power may be divided between eight radiating elements. Thus, 32 transmitter power amplifiers, in all, could be used at one-eighth the power, e.g., 20 watts each, of a single transmitter rated at 160 watts. This feature should be considered minor, however, compared to the complexity required. The antenna system power efficiency is by far the most critical parameter. Attempting to achieve the relative simplicity of the last configuration by increasing the number of elements, and/or arrays to compensate for phase shifter, etc., insertion losses is an un-rewarding trade-off, in that the overall antenna system size becomes quite an installation burden to the ship.

Table 6-4 summarizes a comparison of the various phased-array approaches with that of a switched-beam antenna using a Butler (linear) array, such as illustrated in Figure 5-11. The two antenna categories may be considered to have nominally equivalent gains, depending in each case on switch or phase shifter location. It must generally be concluded that the switched-beam category is more desirable due to higher reliability and less cost.

TABLE 6-4 COMPARISON OF ELECTRONICALLY STEERED ARRAYS AND SWITCHED BEAM ARRAY

		ELECTRONICALLY STEERED ARRAYS			
Array Type	Butler Matrix (Switched) Array	Separate Apertures For Transmit&Receive (Figure 6-3)	Common Apertures 16 Phase Shifters per Array (Figure 6-4)	Common Apertures, 8 Phase Shifters and 8 Diplexers per Array (Figure 6-5)	Common Aperture Single Diplexer (Figure 6-6)
ADVANTAGES	No Phase Shifters: Good Angular Coverage: Reliable due to passive beam-forming network; Lowest Cost	High speed beam scanning; Good Angular Coverage; Graceful Degradation of Array; Commonality of Components			
		Independent Tx and Rx Beam Steering; High Efficiency	Independent Tx and Rx Beam Steering; Moderate Efficiency	Lower Efficiency	Lower Efficiency
DISADVANTAGES	Lossy Feed System: Gain degradation at beam cross-over (even with beam interpolation)	Two Array Apertures: Steering Logic Required: Most Expensive	Steering Logic Required: Very Expensive	Steering Logic Required: Expensive	Steering Logic Required: Less Expensive

The electronically steered category would require slightly larger physical dimensions. The beam steering logic required would virtually be a small special purpose computer. As noted, monopulse-type automatic signal tracking is possible in both the switched-beam and phased array categories, but would be quite complex in terms of RF and receiver circuitry. Accordingly, the slaved stabilization approach, with periodic manual re-alignment, is considered the superior beam pointing/selecting approach for these two categories, at least in the moderate gain ($\leq$ 10 dB) region. Although fan beams have been considered most useful, it would be necessary to slave the beam position control to both azimuth and elevation references, due to the fact that ships' roll dynamics affect relative bearing as well as elevation. (In this context, it should be noted that the preceding comparisons, i.e., Tables 6-3 and 6-4 apply to linear [fan-beam] arrays used in either direction, i.e., horizontal fans as well as vertical fans.) The point is, however, that the requirement for dual-axis slaving requires moderately complex logic to implement the required beam pointing.

In summary, while the concept of electronically-steered phased arrays has the very desirable features associated with frictionless motion, the current complexity of practical implementations is considered herein to negate their recommendation as a moderate gain shipboard antenna for maritime satellite communications.

SECTION 7

SUMMARY AND COMPARISON OF SELECTED APPROACHES

7.1 SUMMARY OF TRADE-OFF ANALYSES

Following a review of the background to maritime mobile satellite services in Section 1, Section 2 of this report described the various performance factors pertinent to and design constraints placed on shipboard antennas intended for widespread use by the civilian maritime community. It was shown that while the large potential maritime mobile user population (many thousands) dictates that the primary link power burden should be placed on the spacecraft, the shipboard terminal should be configured to provide a modest amount of gain. This conclusion is based not only on system capacity considerations and space segment costs, but also on the relationship between shipboard antenna gain and shipboard transmitter power amplifier complexity. The selection of about 10 dB for antenna gain reduces the shipboard transmitter output power requirement to about 100 watts, which is readily achievable with a reliable, low cost transistor array at L-band.

Design Constraints

In general, shipboard antenna gains of 3 to 18 dB are of interest in the context of this study, with 10 dB considered as a baseline reference value for comparison purposes. Section 2 also identified the basic performance parameters of interest in considering antenna designs, all of which are related to

- Minimum Operational Gain
- Manufacturing Cost
- Installation Cost
- Reliability
- Operability

For example, it was shown that considerations of multipath interference and polarization mismatch loss lead to the conclusion that the shipboard antenna should employ circular polarization, a definite constraint to certain types of radiators. It was also shown that the angular dynamics undergone by typical civilian ships are significant, in terms of imposing coverage and rate requirements on antenna design. These are summarized in Table 7-1, along with the above factors.

Three basic categories of antenna subsystem concepts were defined:

1. Mechanically Pointed, Single Beam Antennas
2. Fixed Antennas with Switched Beams
3. Electronically-Steered Phased Arrays

Beam Pointing Technique

Common to all three concepts is the beam pointing problem, i.e., what to use as the basis for beam steering or selection. The consideration of this problem was the primary objective of this study. Accordingly, Section 3 considered the trade offs associated with four different types of antenna pointing:

1. Manual (Operator directed or selected)
2. Slaving (Stabilized to Shipboard Inertial References)
3. Automatic Angle Tracking (of satellite signals)
4. Step-Track (continuous step-scanning)

Manual tracking by operators was shown to be inadequate on its own in terms of system reliability, except when used for very low gain antenna requirements, i.e., 3-5 dB. The manual pointing capability should however be provided in every terminal as a permissible mode of operation, for alignment and initial acquisition.

Antenna pointing by slaving to shipboard inertial references was shown to be a very credible scheme, especially for low-to-moderate gain (< 9 dB) applications. Since virtually all ships of the size considered as potential users are equipped with gyrocompasses, slaving in azimuth would be relatively inexpensive. It was shown, however, that a vertical

TABLE 7-1 SUMMARY OF GENERAL ANTENNA SUBSYSTEM DESIGN CONSTRAINTS

CONSIDERATION	PERFORMANCE REQUIREMENT
System Capacity, Shipboard Transmitter Output Power	Antenna Gain $\approx$ 10 dB
Multipath, Polarization Mismatch	Circular Polarization
Wind Loads	75 Knot Design Criteria
<u>Ship Dynamics</u>: Ship Size >150m	Coverage: 360° Azimuth -30° to +90° Elevation Rates: 15 Deg/Sec, 7.5 Deg/Sec^2
Ship Size <150m	Coverage: 360° Azimuth -45° to + 90° Elevation Rates: 45 Deg/Sec, 45 Deg/Sec^2
Installation Cost	Small Size, Simple Mechanical Interface, No Waveguide. Consider Locations Other Than Mast Head

reference should also be used, since ship rolling in heavy seas can produce large relative azimuth errors as well as elevation errors. This conclusion is true even if vertical fan beam antennas are used. Most ships are not equipped with vertical references; therefore, such must be included in the antenna terminal subsystem when the slaving approach is used. It was shown in Section 3 that a simple inclinometer (pendulum) could serve as an adequate vertical reference, as long as it is located near the roll axis of the ship. The slaved antenna pointing concept is illustrated in Figure 7-1 for the mechanically pointed, single beam antenna category.

The primary disadvantage of the slaved stabilization approach was noted to be the requirement for periodic realignment to compensate for the linear motion of the ship. Depending on the antenna beamwidth and ship speed, the frequency of realignment might be between every 1 and 3 days. The realignment procedure would be manual, accomplished by a crew member steering the beam to the angle of maximum received signal strength. The requirement for periodic manual realignment is a significant disadvantage of the slaving technique, not only because of the burden on the ship's crew, but because of system reliability. This feature could also lower the confidence of the maritime industry and commerce community, in general, in the whole maritime mobile satellite network.

The best alternative to the slaved stabilization approach is automatic angle tracking based on received satellite signal levels. This is conventionally accomplished by using an antenna feed scheme which permits the formation of a null in the (antenna) pattern in the boresight direction. Automatic tracking generally requires a special receiver as well as special antenna feed circuitry. Four basic approaches to automatic tracking are possible:

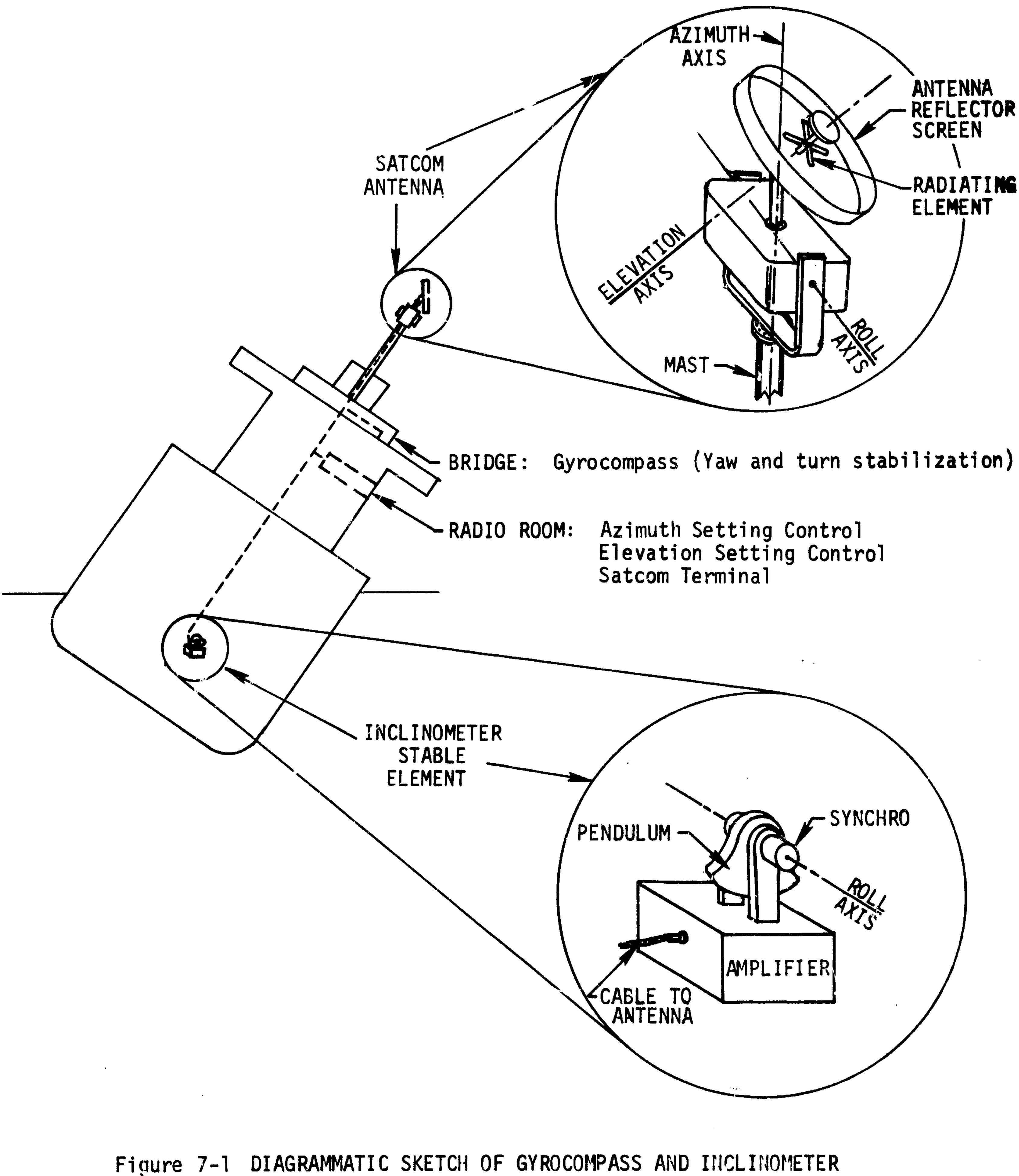


Figure 7-1 DIAGRAMMATIC SKETCH OF GYROCOMPASS AND INCLINOMETER STABILIZED SATCOM ANTENNA SYSTEM

1. Conical-Scanning
2. Three-Channel Monopulse
3. Two-Channel Monopulse
4. Single-Channel Monopulse

Of these, it was shown that single-channel monopulse, alternately referred to as psuedo-conscan or simply monoscan, is the most cost-effective for the subject shipboard application. Multichannel monopulse is more accurate for the same received signal level, but monoscan requires far fewer components and is more reliable. While a special feed circuit is required (a difference channel combiner and a sum channel coupler), no special receiver channel is required. The communications receiver can be used to amplify, transport and translate the tracking error signal to audio, whereupon only special error detectors are required to generate pointing commands. The monoscan concept is illustrated in Figure 7-2, showing the interface with the antenna and communications equipment.

A fourth antenna pointing scheme referred to as step-track was also analyzed. In this concept, no special RF null-forming circuitry is required. The antenna is continually stepped in small increments, and a comparison of received signal levels sequenced in time is made to indicate the direction of maximum level. The technique is essentially a mechanization of manual tracking, and was analyzed to be too rate-limited for shipboard use because of ship motion, as well as requiring relatively high received signal-to-noise ratios.

The conclusion to the analysis of beam pointing or selecting techniques is that both automatic tracking by monoscan and slaved stabilization with periodic realignment are practical. While the equipment costs for the slaving equipment are about half of those for implementing monoscan tracking, the difference between the two is estimated to be

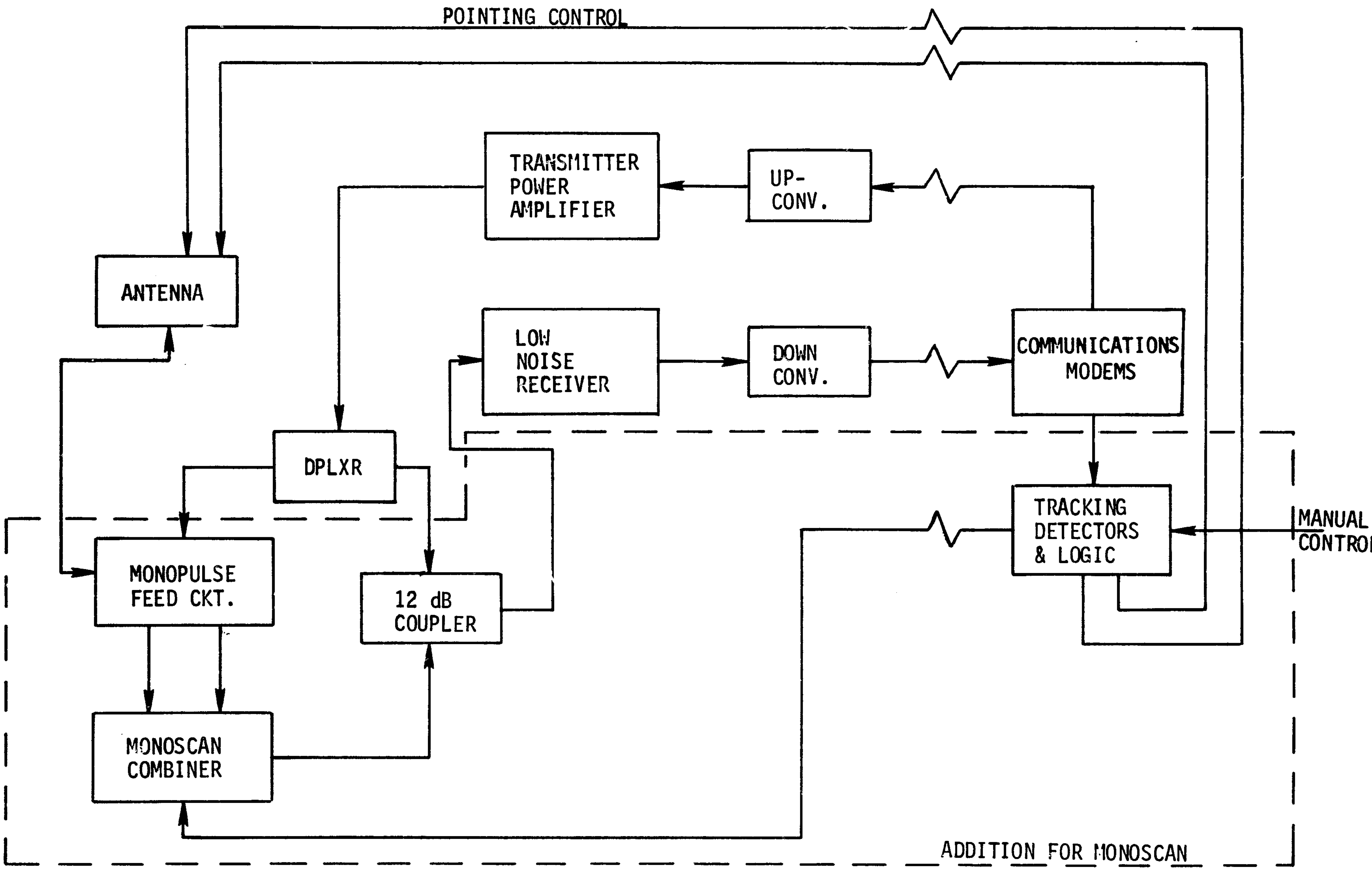


FIGURE 7-2 SHIPBOARD TERMINAL BLOCK DIAGRAM WITH MONOSCAN AUTOTRACKING

reduced to $300 when the additional installation costs associated with inclinometers, reference cables, etc. are included (Table 3-4, page 181). A comparison between the two pointing alternatives also depends on the antenna gain considered. Generally, slaving is more suitable at the lower values of gain, where realignment is required less frequently and autotracking nulls are broader, while autotracking is more effective for higher gain applications. Table 7-2 summarizes these conclusions. In addition, the selection of the best pointing scheme is also dependent on the type of antenna used.

Mechanically-Pointed, Single Beam Antennas

A variety of single beam attenna types suitable for mechanical pointing were reviewed in Section 4. The selection of antenna radiator types is shown in Table 7-3, according to gain requirements. Compared to the cost of pointing mechanisms and circuitry, all these antenna radiators are relatively inexpensive, as indicated in Figure 7-3. For size efficiency and simplicity of manufacture, the short-backfire antenna is of special interest. Its small size can readily provide for a gain of about 14 dBic (dB relative to an isotropic, circularly polarized source), and appears ideal in its physical characteristics. A simple crossed-dipole feed (with 5 or more arms) can be used to provide for a (monopulse) pattern null for monoscan autotracking.

One disadvantage of mechanically-pointed antennas is the need for cable-wraps or rotary joints. Cable wraps are not significant burden by any means, but may not be adequate for the azimuth axis in an elevation-over-azimuth gimbal, since certain shipowners may desire the capability of continuous operation while circling through many revolutions. For this reason, an x-y mount may be desirable. In this arrangement, the two axes of rotation are orthogonal elevation axes. An example of an x-y mount is illustrated in Figure 7-4. Its only disadvantage would be the inconvenience of monitoring, by a shipboard operator, two elevation meters instead of the conventional

TABLE 7-2

BASIC CONCLUSIONS ON BEAM POINTING

TECHNIQUES ACCORDING TO ANTENNA GAIN

Antenna Gain Category	Beam Pointing/Selecting Scheme* Considered Most Cost-Effective
Low Gain Antennas ($\leq$6 dB)	Slaved-Stabilization and/or Manual Tracking
Moderate Gain Antennas (7 - 14 dB)	Monoscan Autotracking or Slaved Stabilization
High Gain Antennas (15 - 18 dB)	Monoscan Autotracking

* As an example, the costs of beam pointing for a mechanically pointed antenna are estimated as follows:

1. Basic Manual Pointing (Pedestal, gearing, remote display, control, etc.) = $2800.

2 a. Additional Cost for Slaving (Dual axis, with peculiar installation costs) = $2200.

2 b. Additional Cost for Monoscan Autotracking = $2500

TABLE 7-3

SELECTION OF MECHANICALLY POINTED

ANTENNAS ACCORDING TO GAIN

Gain Range	Antenna Type	Characteristics
3- 6 dB	Turnstile on Ground Plane	Small and simple, curved for improved axial ratios
7-10 dB	Helix or Horn	Both inexpensive, horn better for tracking
11-15 dB	Short Backfire	Very efficient, simple, fair tracking, small (39 x 10 cm)
16-18 dB	Parabola	Simple and cheap at these gains, but relatively large (60 - 80 cm diameter)

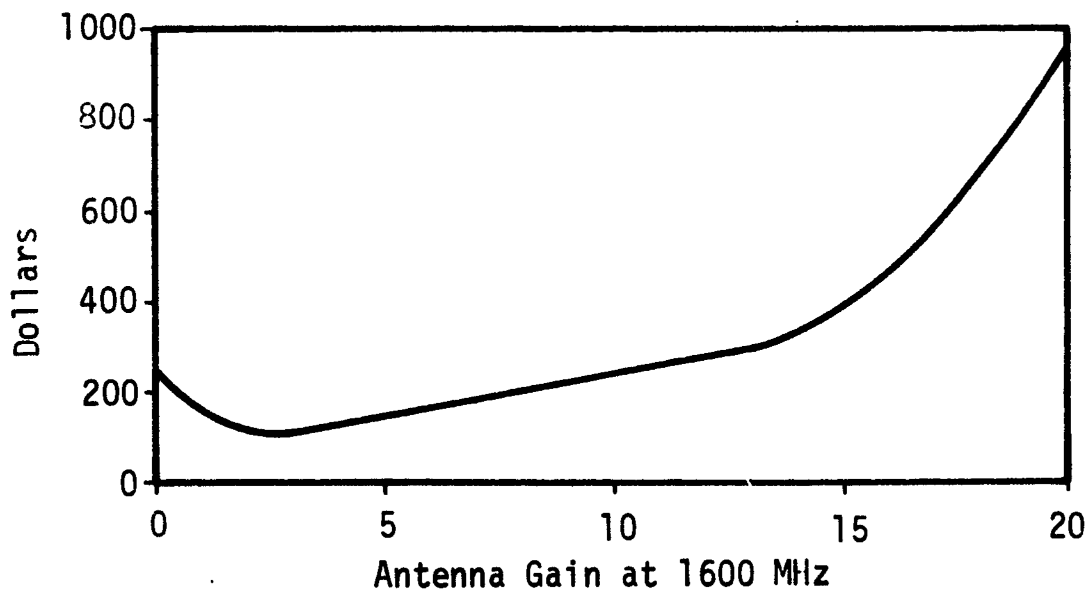


Figure 7-3 BASIC ANTENNA RADIATOR COSTS

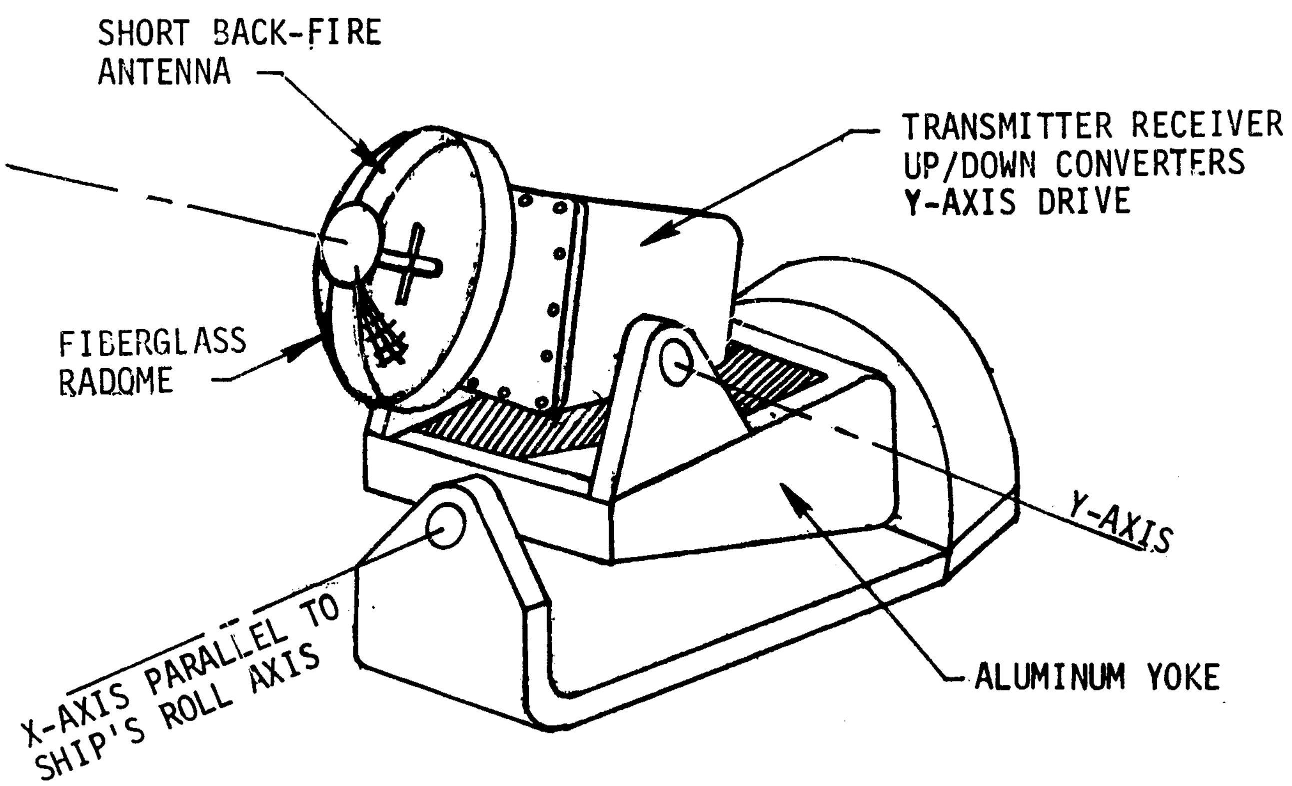


Figure 7-4 CONCEPTUAL X-Y ANTENNA CONFIGURATION

az-el display. This handicap can be overcome, however, either by familiarity or by incorporation of an X-Y to az-el angle data converter.

Switched Beam Antennas

Section 5 addresses a logical alternative to motors, gears and cablewraps, that of switched-beam antennas. In this case, multiple antenna apertures are used, each of which may be attached independently to the ship structure. The locations need not be the mast, but the separation between antennas should not be too great, unless multiple RF transmitters and receivers are used.

Of the example configurations studied, two are considered as representative of attractive schemes. The first is an array of 8 low gain, independent helix antennas, located appropriately about the mast or one of the upper decks of the ship. The second is a set of four Butler-fed linear arrays, each of which produces a fan beam covering about 130^{o} in elevation and 20^{o} in azimuth and steppable through an azimuth range of 90^{o}. A summary of the characteristics of these two antennas is presented in Table 7-4. The use of separate antennas with beam switching is practically limited to low gain applications for the subject shipboard application, due to the volume of equipment resulting from the large number of radiators needed to accommodate the greater-than-hemispherical coverage. The use of fan beams generated by four linear arrays provides for higher gain without excessive size due to the aperture-sharing characteristics of each 8-element array. In order to realize near-full gain of the quadruple array, it is recommended that adjacent beam interpolation circuitry be included. This "fills in the nulls" of the switched-beam array pattern to a large extent (1-2 dB).

TABLE 7-4 PERFORMANCE OF SELECTED SWITCHED BEAM ANTENNAS

Performance Parameter \ Switched-Beam Example	Simple 8-Helix Combination	Quadruple Butler (Linear) 8-Element Array W/ Beam Interpolation
Minimum Gain, dBic	4-5	9-10
Peak Gain, dBic	7-8	10-11
Basic Beam Shape	Circular, 80°	Fan, 20° x 130°
Physical Size	Medium	Large
Complexity Relative Cost	Lowest	Highest
Potential Environment Degradation	Small	Icing Problem

A major trade-off in the configuration of any array is the location of the switch relative to the antenna elements and the transmit-receive diplexer. In order to avoid the degradation of antenna gain and receiver noise temperature implied by the insertion of the beam switch, the number of transmitter power amplifiers and low noise preamplifiers used must be increased from one to the number of elements employed, i.e., 32. For the switched beam antenna category, it was concluded. that the switch insertion loss (perhaps 0.5-1 dB) should be permitted, thereby minimizing RF circuitry.

Electronically Steered Antennas

Finally, Section 6 reviewed the properties of electronically-steered phased array antennas, in terms of their suitability for the shipboard satcom application. The best arrangement would be a set of four linear, 8-element arrays to cover the hemisphere, similar to the higher gain switched-beam example noted above. Given this, major variations in system configuration are possible depending on location of the phase-shifters. As no single array is practical which can accommodate the required coverage, all example configurations are combinations of electronic scanning and beam switching.

Of the various phased array configurations studied, two candidates are most representative of good designs. One is the arrangement, shown in Figure 6-4, of a common aperture with independent steering of transmit and receive beams, and the other is the least complex, with a single diplexer and communications electronics shared between the required arrays.

Of the first, it must be noted that the number of components required is extremely high, although very reasonable efficiencies are achieved by having both the transmit and receive amplifiers between the aperture and the phase shifters. The sheer complexity of the design, when associated with its bulk, means that it would be the least likely candidate to benefit from mass production techniques, and would inevitably be relatively expensive.

The other candidate does not suffer from the same disadvantages as the first, but neither does it have the attendant advantage of high efficiency. The circuitry which is inserted between the aperture and the communications electronics would result in up to 3 or 5 dB of loss, dependent on the scan angle, which will cause a loss of at least 5-6 dB in terminal G/T. This would require four times the satellite power of an equivalent "loss-less" design, or alternatively the array gain, and consequently the number of arrays required, would have to be increased, to the point where the initial advantages of simplicity have been completely negated.

Much of the attenuation (loss) in the approach just discussed is caused by phase shifters. Elimination of these devices by the use of a Butler feed matrix (i.e., switched-beam array) reduces this loss by a considerable amount. The cost is also considerably reduced, since the active diode devices are replaced by passive stripline components which are relatively cheap to produce, especially in quantity. These components are still between the electronics and the aperture, so that the loss is not negligible. Higher gains, and consequently more arrays, are required, to counter this effect and also the gain loss at beam cross-over points. The switched-beam Butler-fed array is still more cost-effective than the electronically-steered array.

7.2 SELECTION AND COMPARISON OF MEDIUM-GAIN CANDIDATE SYSTEMS

As indicated in the preceding summary paragraphs, no single antenna or pointing approach is optimum for all values of required gain between 3 dB and 18 dB. As the nominal baseline gain of 10 dB has been used as a specifically pertinent reference, it is considered appropriate to compare system configurations which provide this basic capability.

One system which must be considered as a very attractive candidate is the single, mechanically pointed short-backfire antenna using single-channel monopulse automatic angle tracking to direct the beam. This is consistent with the conclusions of Sections 3 and 4. The peak gain of this system would be on the order of 14 dB, and the minimum gain realized in operation would be at least 13 dB.

Another system which should be selected is the switched-beam, quadruple linear array which is pointed by slaving to shipboard inertial references. This system is depicted in Figure 5-10. Each of the four arrays is an 8-element Butler-fed linear array producing a fan-beam about 20^{o} wide in azimuth and about 130^{o} in elevation. The peak gain of the array is about 10-11 dB, when the switch insertion loss is included. As this is fairly low, adjacent beam interpolation circuitry should be used to fill in the nulls between beam centers. This would keep the minimum gain at about 9-10 dB as the array is switched in azimuth. As noted in Section 5.3, automatic angle tracking is possible by forming a monopulse pattern, but in this case, the complexity of the RF circuitry could be increased by a significant amount, although similiar circuitry is required for the adjacent beam interpolation function.

It is considered herein that the two different antenna concepts summarized above are representative of practical system solutions for the subject medium-gain, shipboard antennas. As discussed in Section 6, the phased array category is not considered to be competitive with the above two. Table 7-5 presents a general comparison of the two selected approaches.

Examination of the factors listed in Table 7-5 appears to indicate that there is little to compare. The mechanically-driven single-beam antenna is more effective in all aspects. This may be a little misleading. In theory, while it has many more complex RF circuit components, the switched-beam array can have the edge in equipment reliability, since the additional RF circuitry is basically passive. On the other hand, the frequently heard opinion that motor drives are unreliable is repudiated by their overwhelming use. For example, mast-mounted, motor driven X-band radars are virtually standard on all ships.

In essentially all studies of moderate-gain L-band antenna design for aircraft (which the writer is aware of), one form or another of a switched-beam array has been concluded as optimum. Clearly, the aircraft application poses significantly different constraints. The first is the flush-mounting constraint imposed by drag considerations. The second is the coverage required. Civilian aircraft actually do not roll as much or as often as ships. Most aircraft antenna studies have not considered coverage below the (aircraft) horizon necessary. In addition, weight limitations are more stringent and frequent periodic maintenance is more acceptable. Thus, it should not be surprising that a different conclusion exists in the civilian maritime user application.

The costs listed in Table 7-5 should be considered very approximate. While they reflect a certain amount of vendor survey, there is little

TABLE 7-5 COMPARISON OF SELECTED MEDIUM-GAIN ANTENNA SYSTEM CANDIDATES

Parametric Characteristic \ Antenna System	Mechanically Driven Short-Backfire Antenna with Monoscan Tracking	Quadruple 8-element Linear (Butler) Array with Adj Beam Interpolation and Slaved Switching
Peak Gain	14 dB	10-11 dB
Min Operational Gain	13 dB	9-10 dB
System Noise Temp. (Transistor N.F.= 3.5 dB, Preselector Loss=0.5 dB, 15° Satellite Elevation)	525K	580K
Beamwidth	33°	20° x 130°
Axial Ratio for C.P.	Very Good	Fair
Relative Multipath Effects	Normal	Worse
Size, (complete Antenna S/S)	0.35 m^3	1 m^3
Complexity	Minimum	Moderate-High
Periodic Alignment Necessary	No	Yes
Reliability	Good	Good
Potential Environmental Degradation	Minimal	Icing Problem
Manufacturing Cost-Antenna System Incl. RF Head, Pointing S/S	Modest (~7.8K)*	Higher (~10.9K)*
Installation Cost	High (~3-7K)	Higher (~6-14K)

* Assumes at least 200 units per manufacturer

directly applicable vendor experience to warrant confidence in order-of-magnitude estimates of costs of such components in large scale production. A detailed analysis of costs by appropriate component and subsystem manufacturers is beyond the scope of this limited study. Table 7-6 summarizes a breakdown of the equipment costs listed in Table 7-5. They are of course, heavily dependent on many factors, especially quantity of manufacturer. These estimates assume at least 200 units per manufacture. Also, a slightly higher transmitter power is considered appropriate for the switched-beam use. It should be noted that the listed equipment costs do not include the various modems (at least three required) and input-output equipment housed in the radio room.

Installation costs are generally uncertain, and quite variable from ship to ship. They are generally reported to be high for such equipment by shipowners. The switched-beam array installation is considered significantly more tedious, since there are more units to install and careful alignment of each is required during installation. In addition, the dual axis slaving approach adds to the installation burden.

A final point to be made is with respect to mast location and system redundancy. While the top of the mast is by far the most desirable location for the selected mechanically-pointed antenna system, certain shipowners may insist it be placed elsewhere, simply because of prior crowding. If no other superstructure location exists which provides for the required greater-than-hemispherical coverage, then two antenas may have to be used, mounted for example on opposite sides of an upper deck. Such a location would be more of a burden on the mechanically-pointed, single beam antenna than on the switched beam array, which is readily amenable to such a separation. It would add about $2500 to the cost of equipment in the mechanically-pointed case, relative to zero in the array configuration. This brings the equipment

TABLE 7-6

COST ESTIMATE BREAKDOWN OF ANTENNA SUBSYSTEM
EQUIPMENT MANUFACTURE FOR SELECTED DESIGNS

Mechanically Pointed, Autotracking Antenna	
Short-Backfire Antenna, with Radome	$ 300
Pedestal, Motors, Display, Controller	2,200*
Monoscan Equipment, incl. Feed.	2,500*
Diplexer, RF Lines	500
Low Noise Preamplifier	300
Transmitter, <100 w	1,200
Up & Down Converters	600
Cabling, Miscellaneous	200
Total	$ 7,800
Switched-Beam Linear Array	
4 Linear Arrays, 8 Elements each	$ 3,000
Butler Feeds (4) with Beam Interpolation	2,800
Slaving Equipment, Dual Axis	1,200*
Switching Logic	800
Diplexer, RF Lines	500
Low Noise Preamplifier	300
Transmitter, >100 w	1,400
Up & Down Converters	600
Cabling, Miscellaneous	300
Total	$10,900

* See Table 3-4 for further breakdown.

costs much closer together. The high gain of the short-backfire antenna makes it more amenable to use of a single RF head (diplexer, transmitter power amplifier, low noise preamplifier and up/down converters) between the two antennas, however. On the other hand, dual RF heads could be used with both system concepts for about an additional \$2500 each. The obvious redundancy of such a dual antenna installation would have certain advantages in communications reliability. An automatic handover switching function could have to be implemented, and this would be an additional few hundred dollars. Thus, complete RF redundancy may be provided for about \$5.5k with the mechanically-pointed system and for about \$2.9k in the switched-beam array system.

Thus, the general conclusion of the comparison is that the mechanically-pointed short-backfire antenna subsystem is the better design approach for the majority of users. Its characteristics and performance are summarized in Table 7-7. It should be noted that the estimated cost of \$7800 in Table 7-6 for the basic antenna subsystem includes antenna mounted RF electronics, but does not include the costs of the basic radio equipment needed in the radio room, or general installation. In addition to the low frequency circuits for antenna position control, the radio room electronics include IF distribution (70 MHz) circuitry, an access control modem, a voice modem, a data/telex modem, a teletype and a small console for housing. This equipment is estimated to cost an additional \$4700. Thus the total cost of the terminal equipment is estimated to be about \$12,500. For ease of reference and comparison in context, a breakdown of the total shipboard system cost is presented in Table 7-8. In many cases, there could be an additional 15% in price for vendor mark-up. As noted, installation costs can vary a great deal, but typically might average between \$4000 and \$5000 for the complete shipboard terminal.

TABLE 7-7

SUMMARY OF CHARACTERISTICS AND PERFORMANCE
OF RECOMMENDED SHIPBOARD ANTENNA SUBSYSTEM

Antenna Subsystem Characteristic	Performance
Antenna Gain, 1535-1543.5 MHz	14.0 dB peak, (13.0 dB min. operating in system)
Antenna Gain, 1636.5-1645 MHz	14.2 dB peak
Antenna Radiator Type	Short-Backfire
Antenna Feed	Monopulse, (>0.02 volts/volt-deg)
Antenna Size	39 cm (15.4 in.) diam. by 10 cm (3.9 in.) deep
Half-Power Beamwidth	~33°
Polarization	Circular
Pedestal	X-Y Mount
Coverage:	360° Azimuth, and:
Ship size >150 m	-30° to 90° elevation
Ship size <150 m	-45° to 90° elevation
Angular Rate Capability:	
Ship size >150 m	15 deg/sec, 7.5 deg/sec^2
Ship size <150 m	45 deg/sec, 45 deg/sec^2
Wind Loads	75 knot design criteria, Survivability at 150 knots
Pointing Modes:	
1) Manual	Console operator
2) Autotracking	Single-channel Monopulse (Monoscan)
Monoscan Coupling Factor	12 dB
Position Tkg Loop (Noise) Bandwidth	2 Hz (Large Ships), 4 Hz (Small ships)
Electronics Mounted With Antenna	Diplexer, Preamplifier (3.5 dB N.F.) Solid State Transmitter (~60 W), L-band/70 MHz Up- and Down-Converters

TABLE 7-8

ESTIMATE OF EQUIPMENT COSTS FOR COMPLETE
SHIPBOARD TERMINAL SYSTEM

Subsystem/Component	Est. Cost
Short-Backfire Antenna, with Radome	$ 300
Pedestal, Motors, Display, Controller, etc.	2,200
Autotracking Electronics, incl. Feed and Servo	2,500
Diplexer, RF Lines	500
Low noise Preamplifier (Transistor)	300
Transmitter ($<$100 w)	1,200
Up- and Down-Converters	600
Cabling (70 MHz Comm)	200
IF Distribution (70 MHz)	300
Access Control Modem	900
Voice Modem	800
Data and Telex Modem	1,200
Input/Output Device (TTY)	1,400
Rack/Console	100
Total Terminal	$12,500

SECTION 8

CONCLUSIONS AND RECOMMENDATIONS

8.1 RE-STATEMENT OF OBJECTIVES

As this study is the first known to specifically address L-band shipboard user antennas for use with a civilian maritime services satellite system, it was intended to be parametric in nature rather than directed at a specific antenna proposal. The fundamental objectives of the study were to:

1) Perform and document a basic conceptual investigation of shipboard antenna subsystems which would operate in the band 1535 MHz to 1645 MHz, with

2) Gain considered as a study parameter over the general range 3 dB to 18 dB, by

3) Identifying the various pertinent performance characteristics and requirements and related design trade-offs, and by

4) Comparing alternate approaches for their simplicity and general suitability.

The study, thus directed, is intended for use by system planners in postulating maritime satellite system configurations, link sizing and economic trade-offs.

As an ancillary result from the study, a specific antenna configuration emerged as most cost-effective. Thus, additional conclusions and recommendations relating to this antenna are also presented in the following pages.

8.2 CONCLUSIONS

8.2.1 Technical Conclusions

Previous analyses of performance requirements indicated that the shipboard antenna should have a gain, in the receive band of 1535 MHz to 1543.5 MHz, on the order of 10 dB or more, assuming an earth-coverage spacecraft antenna and a capacity equivalent to 10 voice channels. This results in a requirement for less than 100 watts of shipboard RF transmitter power, which is relatively inexpensive.

Consistent with this basic sizing, the most fundamental conclusion to this study is that shipboard antenna subsystems providing a moderate amount of gain, i.e., 9-14 dB, are in fact feasible, practical and economically viable within the concept of a commercial maritime services satellite system.

Specifically, based on an analysis of environmental constraints and performance versus complexity comparisons of various pointing and antenna techniques, the following antenna subsystem design approach was concluded as singularly optimum:

Basic approach:	Mechanically pointed, single beam antenna on a small, dual-axis mount
Antenna type:	Short-backfire radiator, 39 cm (15.4 inch) diameter by 10 cm (3.9 inch) deep, with ≥14 dB peak gain
Antenna feed:	Monopulse (crossed-dipoles), circular polarization
Antenna pointing:	Automatic angle tracking of satellite using single-channel monopulse circuitry with manual control for back-up and acquisition

Minimum gain:	13 dB in operation (with received signal level of 45 dB-Hz)
Spatial coverage:	360° azimuth, -45° to 90° elevation for ships <150 meters in length; -30° to 90° elevation for larger ships
Tracking rates:	45 deg/sec, 45 deg/sec^2 for ships <150 meters; 15 deg/sec, 7.5 deg/sec^2 for larger ships
Antenna location:	Mast-mounted wherever possible
Antenna/mount subsystem weight:	<50 lbs, including RF electronics (diplexer, low noise preamplifier, 60 watt transmitter, up- and down-converters)

The mechanically-pointed, single beam antenna approach was concluded to be superior to the alternate categories of multibeam antenna arrays with beam switching and electronically-steered phased arrays for the following reasons:

- Lower total equipment cost (for gains >4 to 5 dB)
- Smaller overall size (swept volume) for comparable gain
- Relative ease of achieving the greater-than-hemispherical coverage required
- Better overall performance (gain, noise temperature, circular polarization, pointing error)
- More self-contained and thus lower installation costs
- Amenability to use of single RF transmitter/preamplifier at antenna

For pointing of the directive beam (about 33° beamwidth for the short-backfire antenna), it was concluded that autotracking of the satellite signal is the best approach. While manual pointing by the radio operator would certainly be less expensive, such an approach was concluded to be adequate only for very low gain applications (i.e., 3 or 4 dB), wherein fairly infrequent switching would be possible. This inadequancy of manual pointing is due to the high relative angular dynamics attendant with ships motion in heavy seas. A manual remote position control mode is recommended, however, as a back-up and for initial acquisition.

Slaving of the antenna beam to shipboard inertial references (the gyrocompass and an inclinometer) was concluded to be an adequate solution for antennas of moderate gain (9-14 dB); however, it was negated in favor of satellite signal autotracking for the following reasons:

- An analysis of ships motion showed that large relative azimuth errors due to ships rolling can occur (when the satellite is in the direction of the bow-stern line), and thus making necessary the installation of a special inclinometer (near the ships roll axis)
- Lower pointing error, for roughly equivalent costs when extra installation costs associated with gyrocompass and inclinometer slaving are included.
- Higher reliability
- Better self-containment; one interfacing cable route instead of three
- No periodic manual realignment required to compensate for ships daily headway

Of the possible approaches to automatic satellite signal tracking (multi-channel and single-channel monopulse, conical-scan, Step-Track), it was concluded that single-channel monopulse (monoscan) is the optimum choice for the shipboard application because

- Monoscan angle tracking performance (sensitivity, pointing error) is adequate
- The Step-Track technique is not feasible due to the combination of high angular dynamics and low received signal levels, and thus
- Monoscan is the simplest, least complex and least expensive of the usable techniques
- Compared to the more accurate multi-channel monopulse schemes, monoscan requires no additional RF or IF receiver channels and thus no periodic intra-channel phase and gain alignments
- Because of the above, the monoscan approach has the highest operational reliability

The complete shipboard terminal, concluded as being most cost-effective, consists of three basic entities:

1) Antenna Assembly - This includes the short-backfire antenna covered with fiberglass, pedestal and motors, diplexer, monoscan combiner and RF portion of the radio electronics. The 60 watt transmitter, low noise preamplifier and up- and down-converters (filter, mixer, RF multiplier and IF amplifier/driver portions) are recommended for housing at the antenna for the obvious reasons of achieving the maximum efficiency attendant with negligible RF line loss, and the relative ease of long cable run interface at IF. It is estimated that the total weight of the antenna assembly should be less than 50 lbs. For complete coverage, it is clear that the most

effective location for the antenna assembly is as close to the mast as possible - ideally above the radar antenna(s).

2) Interface - Because the RF electronics are antenna mounted, the cabling between the antenna assembly and the radio room can be a harness of relatively small diameter, flexible, coax. This interface consists of two local oscillator signals (typically fixed in the 40-50 MHz region), the up- and down-link communication signals at IF (e.g., 70 MHz) the monoscan timing signal, antenna motor control signals (two) and ac and dc power. (No power supplies would be included in the antenna RF package.)

3) Radio Room Electronics - A console would be installed below deck (in the radio room) which need consist only of four small functional chassis:

- Access Control Modem (including IF distribution)
- Voice Modem
- Data/Telex Modem
- Antenna Controlle (including disp ld tracking circuitry)

As these chassis would be small, the uld be housed in less than one-half a standard rack - or preferably in two table mounted consolettes. In addition, a conventional teletype would be required.

For ships severely constrained by overcrowding of equipment on masts and/or other suitable superstructures, wherein installation of dual antenna subsystems on bridge decks is considered, the cost-effectiveness superiority of the mechanically-pointed short-backfire antenna over an assembly of four linear, 8-element arrays designed for switching of vertical fan beams about the periphery of the ship, is estimated to be reduced to a small amount. Even in this case, however, the superiority of the mechanically pointed single beam antenna approach is predicted to exist.

Although a single antenna subsystem design approach has been concluded herein as superior, it must be noted that it is by no means necessary that all ships using the satellite system be equipped with precisely the same type of terminal. Clearly, different terminal designs would be possible in the system, as long as certain basic specifications, e.g., shipboard EIRP (effective isotropic radiated power) and receive G/T (antenna gain to receiving system noise temperature ratio), modulation character and formats, modem performance, selectivity and carrier stability were met.

8.2.2 Economic Conclusions

A basic conclusion to this study is that the cost of providing a moderate amount of antenna gain in the shipboard terminal is essentially the cost associated with the equipment required for the beam pointing function. This is especially evident in the mechanically-pointed, single beam antenna subsystem concluded as the best approach. This and related pertinent conclusions are highlighted below:

1. The Cost of Beam Pointing

   - Excluding installation and the antenna radiator itself, the cost of beam pointing, including feed, pedestal and single channel monopulse (monoscan) autotracking receiver circuitry, etc. is estimated to be about \$4700. (See Table 8-1 for breakdown.)
   - The cost of the antenna radiator itself is relatively small, being on the order of \$200-\$300 for the ranges of 0-1 dB and about 6-14 dB, and less than \$200 between

TABLE 8-1

COST ESTIMATE OF BEAM POINTING EQUIPMENT
FOR MECHANICALLY DRIVEN (SHORT BACKFIRE)
ANTENNA WITH MONOSCAN AUTOTRACKING
(assuming 200 unit production base)

Subsystem/Component	Cost
1. Basic Manual Pointable Mechanism:	
Antenna Pedestal, Gearing	$ 500
Drive Motors	400
Control Lines	100
Position Display, Manual Controller	900
Miscellaneous Furnishings	300
Subtotal, Basic Pointing	$2,200
2. Monoscan Tracking Electronics:	
Monopulse Feed and Comparator	$ 950
Monoscan Converter and Coupler	850
Scan Generator	100
Sync. Angle Detectors, Filters	200
Servo Position Loop Amplifier	400
Subtotal, Monoscan	$2,500
Total, Tracking Subsystem	$4,700

1 and 6 dB. Fiberglass radomes and foam encapsulation are included in these estimates.

- The $4700 tracking equipment cost assumed a 14 dB peak gain, short-backfire antenna. For a 0 to 1 dB antenna, there would be no pointing cost, while for up to a 5 to

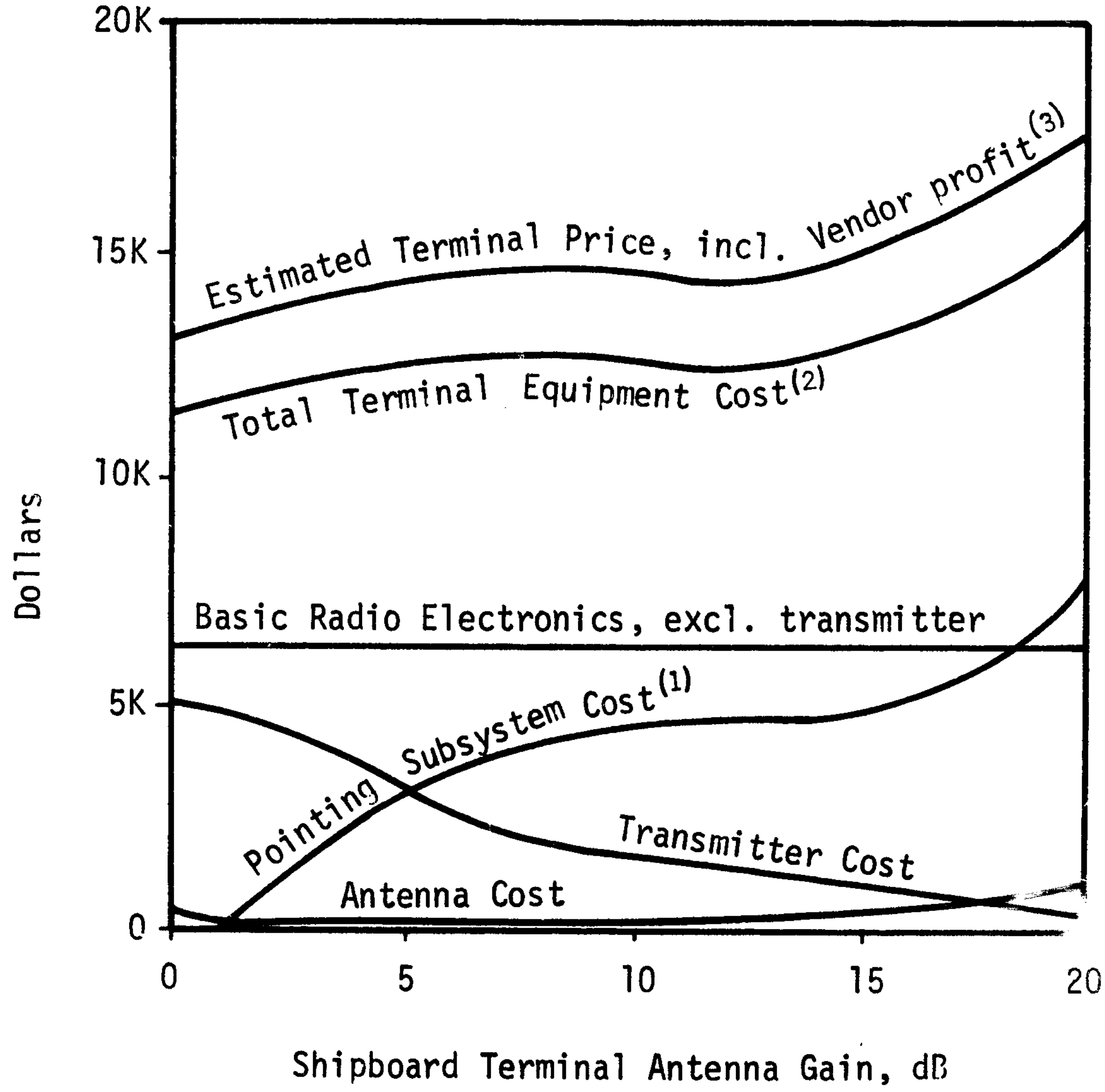
6 dB gain antenna, modest cost azimuth-only slaving might suffice. For high-gain antennas (e.g., >18 dB), the pedestal and drive motor costs increase significantly.

2. Shipboard Transmitter Costs

- A major advantage of providing a moderate amount of antenna gain onboard the ship is that it reduces the transmitter power requirement. For the $5000 pedestal, antenna and tracking system described in item 1 above, approximately a 60 watt transmitter costing less than $1200 is adequate.
- For comparison, from Table 7-8, the total shipboard terminal cost including all baseband (voice, data, access control) modems and an input/output device was shown to be $12,500. Thus, the allocation of approximately 40% of the terminal cost to the beam pointing function permits use of a transmitter costing only 10%. By contrast, a hemispherical antenna costing $200-$300, and necessitating no pointing, would require about a 1 kW transmitter employing a traveling wave tube (TWT) amplifier and costing about $5000. Thus, the cost of the beam pointing function tends largely to be saved in reduced transmitter costs. More importantly, the higher the shipboard antenna gain, the lower the required satellite transmitter power and hence space segment cost (i.e., beam pointing is cost-effective at both ends of the communication link).

3. Shipboard Terminal Cost Vs Performance Trade-Offs

- The shipboard terminal cost vs performance trade-offs are illustrated in Figure 8-1 as a function of shipboard antenna (receive) gain, where terminal system cost is



Notes:

1. Pointing subsystem costs include all appropriate mechanical and electronic components necessary to function.
2. All equipment costs assume manufacture of at least 200 units.
3. Estimate Vendor Price includes 15% mark-up.

Figure 8-1 ESTIMATE OF SHIPBOARD TERMINAL SYSTEM COSTS (Excluding Installation) constituted by antenna radiator cost, pointing subsystem cost, transmitter subsystem cost and the cost of the basic remaining radio electronics, which is constant at $6,300. A vendor or manufacturer's representative mark-up of 15% is included in the estimated price to the shipowner.

- It is concluded, therefore, that the terminal system price to the shipowner may, in fact, be almost constant, at about $14,000 up to about 13-14 dB, where it is $14,375 for the recommended autotracking short-backfire antenna approach. Above this value of gain, not only do equipment costs increase significantly, but so do installation costs.

4. <u>Shipboard Installation Costs</u>

- It is estimated that when the electronics console and cabling are included, the cost of installation of a satellite terminal on a commercial maritime vessel would be about $3,000 even with a simple, small, fixed, hemispherical (1 dB gain) antenna.
- For moderate gain, modest size antennas, such as the short-backfire with a small gimbal, the added welding and brackets for superstructure mounting are estimated to increase the costs by over $1,000.
- As the antenna and pedestal system becomes larger at higher gains (e.g.,>18 dB), a crane must be used as well as increased support structure sizes, such as platforms, so that the installation task could run $7,000 to $8,000.
- Figure 8-2 shows the effect of these installation costs on the total cost of the installed satellite terminal system to the shipowner.

On this basis then, it must be concluded that an antenna gain of 13 dB net is, in fact, a cost effective selection, given that some gain above 3 dB is required. Above about 14 dB, it can be expected that costs will

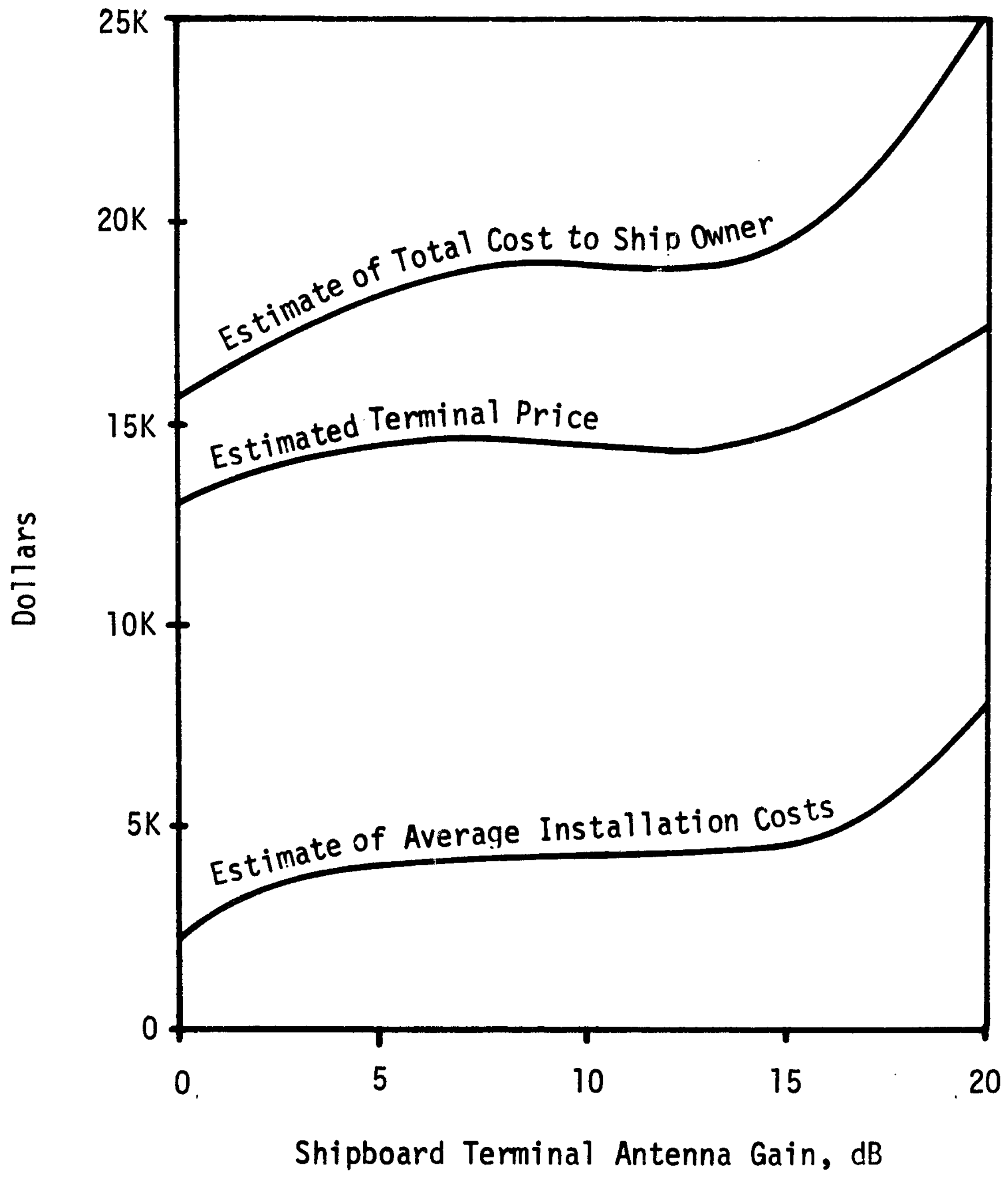


Figure 8-2 ESTIMATE OF TOTAL SATELLITE TERMINAL COST TO SHIP OWNER INCLUDING INSTALLATION

increase at a very rapid rate, given that the antenna is designed to meet the same angular rate, wind speed and coverage requirement. The high costs of shipboard terminals indicated in Figures 8-1 and 8-2 for the very low gain antennas (1-4 dB) are due to the increased shipboard transmitter power requirements, and are thereby based on the assumption that the spacecraft antenna has earth coverage in its receive band.

The results of Figure 8-2 can be used to estimate the total investment by maritime satellite users in shipboard terminals, according to antenna gain and user population size. For example, for a user population of 10,000 ships, the difference between 3 dB of antenna gain and 13 dB of antenna gain is less than $20 million. This would appear to be a small price to pay for ten times the satellite capacity.

Thus, the conclusion may be made that the shipboard terminals should have a moderate amount of antenna gain; and given this, that the mechanically-driven short-backfire antenna with automatic angle tracking, providing a minimum gain of about 13 dB, is most effective.

8.3 RECOMMENDATIONS

Two basic recommendations are offered as a consequence of this antenna study:

1. It is suggested that ship operators, both commercial and government, consider the mechanically-pointed, short back-fire antenna subsystem for its favorable features and potential benefits for maritime satellite communications.
2. It is recommended that a development program be initiated for a few prototype terminals based on the design concepts presented in this study, so that meaningful technology demonstrations and sea trials can be conducted with forthcoming experimental and/or preoperational L-band spacecraft (e.g., ATS-F, and/or possibly the proposed DOT or DOT/ESRO Aerosat, etc.)

Consistent with recommendation number 2 above, it is believed that little, if any, further system analyses or trade-off studies are required to develop a credible shipboard terminal. It is vital, however, that certain critical components of the recommended antenna subsystem be placed under development as soon as possible. (In this regard, it should be noted that no applicable antenna subsystem experimentation has been conducted to date, and that most of the required components are not readily available.) A certain amount of specific design analysis in conjunction with the prototype hardware development will be necessary. The critical development items for the recommended user antenna subsystem include the following:

1. Development of a 1535 MHz/1645 MHz short-backfire antenna with monopulse feed. (A multiple dipole feed - number of arms ≥5 - is suggested.)
2. Design and development of small 1540 MHz monopulse comparator

(e.g., in stripline).

3. Design and development of small, lightweight, low-loss, inexpensive narrowband monoscan combiner and coupler (possibly integral with comparator).

4. Design and development of small, lightweight, dual-axis pedestal with drive motors for the 39 cm (15.4 inch) diameter antenna, about 20 pound-feet torque, and angle data pick-offs with 1° resolution (a dual-elevation, i.e., X-Y mount with cable wrap appears preferable; however, further consideration should be given to an elevation over azimuth mount).

5. Design and development of antenna position controller chassis, to include remote control electronics, position display, manual/automatic mode switch, servo amplifiers, monoscan timing generator, synchronous (angle) phase detectors and filters (to condition inputs from communication receiver AGC).

6. Integration of Items 1-5 with radio electronics.

While not strictly a functional part of the antenna subsystem, the RF electronics package to be mounted with, or possibly adjacent to, the antenna should also be developed. It is especially important that this assembly be mechanically designed in conjunction with the antenna gimbal, Item 4 above. In order to achieve this necessary integration, the following are required:

7. Design and construction of a simple 1.54 GHz/1.64GHz diplexer.

8. Design and development of a small and light, solid-state, 60 watt, 1.64 GHz transmitter.

9. Procurement of a 1.54 GHZ preselector filter and low noise, transistor preamplifier.

10. Design and development of up- and down-converters, including mixers for an IF of 70 MHz, filters, L.O. multipliers and driver amplifiers (excluding, but specifying, local oscillator input frequencies and stabilities to be externally provided).

11. Mechanical design and fabrication of a compact weatherproof enclosure for above RF electronics (in conjunction with antenna gimbal design effort).

Finally, an antenna subsystem by itself is insufficient for the intended sea trials. Modems should be obtained which are representative of those considered appropriate for operational use. In addition to antenna subsystem performance demonstrations (autotracking, voice and data quality) a very critical aspect of maritime satellite services will be system access. An access control technique (such as that recommended in previous AMI studies) should be selected and a corresponding access control modem developed which is representative of an approach that permits efficient access of a few channels by thousands of ships, in both routine and emergency modes.

It is estimated that it would take the better part of one year to develop the subsystem components delineated above and assemble the prototype terminal(s). With NASA's ATS-F spacecraft soon to be available, and/or possibly the proposed DOT or DOT/ESRO initial Aeronautical Satellite, and/or other NASA vehicles, it is recommended that the development effort summarized above be initiated in the very near future.

While it is believed that little further system trade-offs are necessary to the commencement of the development of the terminal recommended herein, it should be noted that one major maritime satellite service - navigation -

has not been considered to any great extent in this study. Eventually, the maritime satellite services will include navigation capabilities, and if, as considered probable, the prescribed navigation technique incorporates multiple synchronous satellites, then an additional major requirement could be placed on the shipboard user terminal. It is accordingly recommended that further study of satellite navigation aspects be made specifically relative to shipboard terminal configurations.

Given the probability of the multiple satellite approach (each of which would also be communication relays), it may be practical to take advantage of the fact that the signal energy requirements for navigation services should be quite small relative to the communication services. This difference could be such that two independent high gain shipboard antennas would not be needed. Given an analysis of satellite separation versus signal energy requirements, it is conceivable that a single beam is adequate; or that a separate, small, non-transmitting or non-receiving, fixed hemispherical antenna could be used to provide the additional range vector. Clearly, this subject warrants further study. Such an analysis, however, should not impede the recommended shipboard antenna terminal development.

APPENDICES

APPENDIX A

ANGLE TRACKING ERROR DUE TO THERMAL NOISE IN A MONOPULSE RECEIVER

Figure A-1 depicts a simplified block diagram of a monopulse angle tracking system. The scheme characteristically employs three separate receiver channels - a sum channel and two difference channels. The channels are separately but dependently down-converted from RF to an intermediate frequency (IF) where pointing error signals are generated by product detectors. The error signals are then used to drive the servomechanism which in turn maintains the antenna pointing toward the signal source. In this manner, the autotracking loop is closed. The pointing mechanism may be an antenna drive motor, a set of beam-switches or phase-shifters in a phased array.

The use of the product detectors in the receiver to form the product of the sum and difference channels constitutes cross-correlation of the monopulse sum and difference signals, and as such characterizes the tracking scheme. When the RF signal permits (e.g., a CW carrier), a phase-locked detector may be used in the sum channel, and separate phase detectors employed in the difference channels which in turn utilize the sum channel local oscillators (and phase-locked VCO) as references. In this latter receiver configuration, the basic cross-correlation model still applies.

While there are numerous causes of pointing error in such a tracking scheme such as wind gusts, dynamic lag, amplitude and phase unbalance between channels, etc., the most fundamental limitation in tracking performance is that due to thermal noise in the receiver. The following paragraphs summarize the derivation of the tracking error due to thermal

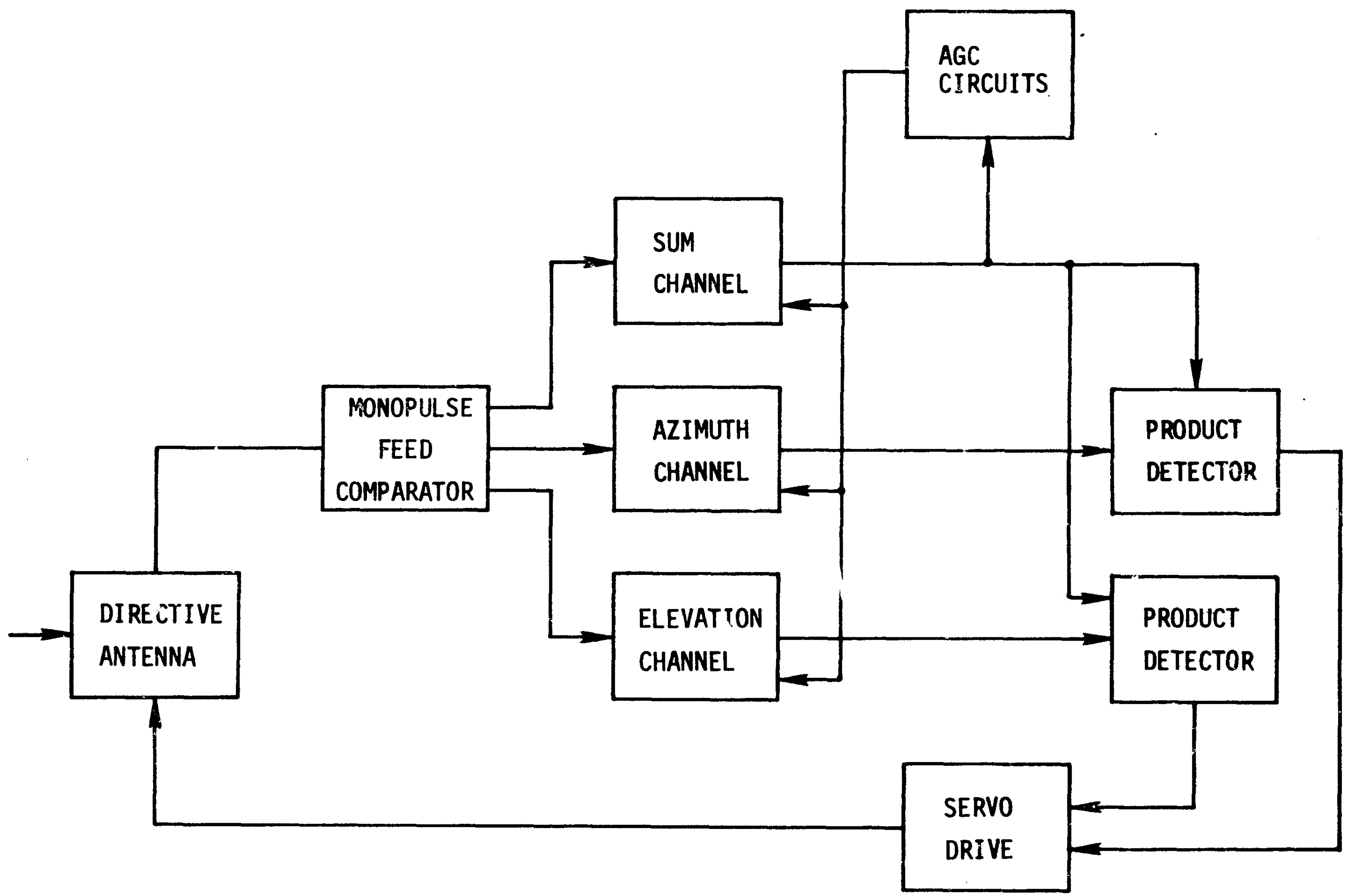


Figure A-1 MONOPULSE ANGLE TRACKING

noise in terms of the design parameters of a monopulse angle tracking system. These parameters include the monopulse feed (pattern) characteristic, the receiver noise figures and bandwidths and the servo bandwidth, as well as received signal levels.

For the purpose of analyzing the performance of the monopulse tracking technique, a model for the error detection portion of the system is assumed and is shown in Figure A-2. The symbols used in this diagram are defined as follows:

$G_\Sigma(\Theta)$ = voltage gain of the sum channel at an angle Θ off boresight

$G_\Delta(\Theta)$ = voltage gain of the difference channel at an angle Θ off boresight

Θ = the angle between boresight and the direction of the RF signal

$S(t)$ = signal at the antenna aperture

$n_\Sigma(t)$ = sum channel noise

$n_\Delta(t)$ = difference channel noise

$n_p(t)$ = product detector noise

The inputs to the product detector may be written as:

$$u(t) = G_A[G_\Sigma(\Theta)\ S(t) + n_\Sigma(t)] \tag{1}$$

$$v(t) = kG_A[G_\Delta(\Theta)\ S(t) + n_\Delta(t)] \tag{2}$$

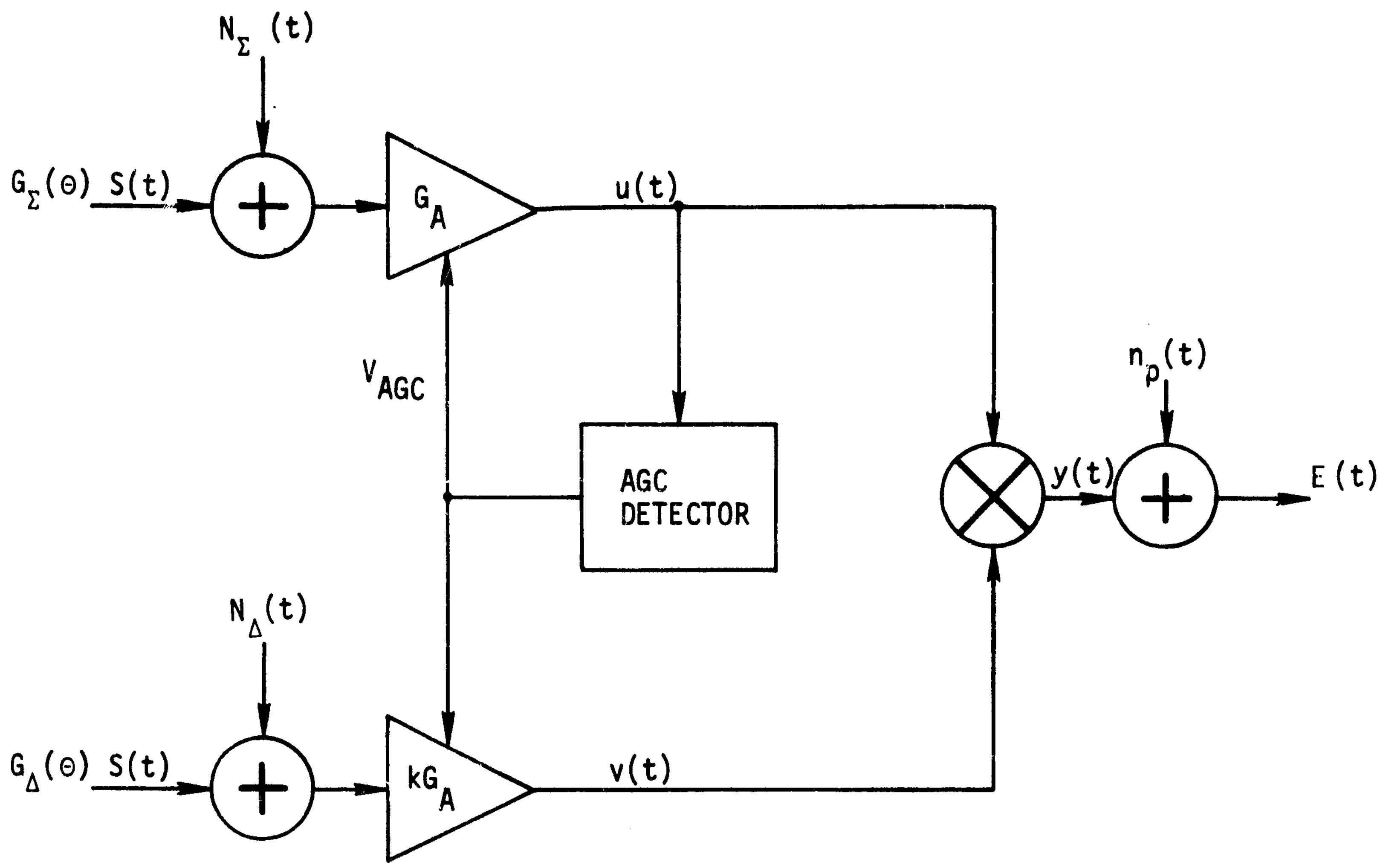


Figure A-2 MONOPULSE RECEIVER ERROR DETECTION MODEL where

G_A = AGC amplifier gain

k = error channel gain proportionality constant

The product detector output then becomes:

$$E(t) = akG_A^2 \left[G_\Sigma(\Theta)\, G_\Delta(\Theta)\, S^2(t) + G_\Sigma(\Theta)\, S(t)\, n_\Delta(t) + G_\Delta(\Theta)\, S(t)\, n_\Sigma(t) + n_\Sigma(t)\, n_\Delta(t)\right] + n_p(t) \quad (3)$$

where

a = product detector gain constant.

The first term of Equation (3) contains the desired error information and the noise is represented by the remaining terms.

Assuming that $S(t)$, $n_\Sigma(t)$, $n_\Delta(t)$ and $n_p(t)$ are uncorrelated, the power spectrum of $E(t)$ is given by:

$$S_E(f) = a^2k^2G_A^4 \left[G_\Sigma^2(\Theta)\, G_\Delta^2(\Theta)\, S_{s^2}(f) + G_\Sigma^2(\Theta)\, S_s(f) \times S_{n\Delta}(f) + G_\Delta^2(\Theta)\, S_s(f) \times S_{n\Sigma}(f) + S_{n\Sigma}(f) \times S_{n\Delta}(f)\right] + S_{np}(f) \quad (4)$$

Equation (4) indicates that by increasing the error channel gain constant, k, the effect of the product detector noise can be minimized. The upper limit of this gain constant, k, is restricted by the saturation characteristics of the product detector. In general, the amplifier gain is adjusted for zero error voltage in the normal tracking condition, i.e., in the on-boresight condition. Prior to equating the desired (first) term in and noise (remaining) terms in (4), certain assumptions are warranted for tractability of the result. These are the following:

a) The AGC in the sum channel maintains the amplified signal output at a constant value.

b) The product detector (self) noise $n_p(t)$ is negligible.

c) The tracking error is small, so that $G_\Delta(\Theta)\ S(t)\ n_\Sigma(t)$ is negligible and $G_\Sigma(\Theta) = G_\Sigma(0)$.

d) The difference channel voltage gain $G_\Delta(\Theta)$ may be expressed as

$$G_\Delta(\Theta) = k_\Delta\ G_\Sigma(\Theta)\ \Theta = k_\Delta\ G_\Sigma(0)\ \Theta \tag{5}$$

where k_Δ is the slope of the difference channel error signal (voltage) pattern normalized (via AGC) to the sum channel, in volts/volt-degree. (See Appendix B).

With the above assumptions, the rms angle tracking error (standard deviation in Θ) may be reduced from Equation (4) and expressed for convenience as

$$\sigma_{\Theta n}^2 = \frac{1}{k_\Delta^2} \frac{\left(1 + \frac{S}{N_\Sigma}\right)}{\left(\frac{S}{N_\Sigma}\right)\left(\frac{S}{N_\Delta}\right)} \frac{B_\Delta}{B_{IF}} \tag{6}$$

In this expression,

S = signal power (sum channel reference)

N_Σ = sum channel noise power in IF bandwidth

N_Δ = difference channel noise power in IF bandwidth

B_{IF} = IF bandwidth

B_Δ = servo position loop (two-sided) noise bandwidth

For the case of relatively high signal-to-noise ratio ($S/N_\Sigma >> 1$) in the sum channel (such as would be typical in a phase locked loop, wherein the loop bandwidth is used instead of IF bandwidth) and upon noting that $N_\Sigma = k\, T_\Delta\, B_{IF}$, where T_Δ is the receiving system noise temperature of the difference channel(s) and k is Boltzmann's constant,

$$\sigma^2_{\theta n} = \frac{1}{k^2_\Delta} \frac{N_\Delta}{S} \frac{B_\Delta}{B_{IF}} = \frac{1}{k^2_\Delta}\left(\frac{k\, T_\Delta\, B_\Delta}{S}\right) \qquad (7)$$

Another equivalent expression which is frequently convenient is

$$\sigma^2_{\theta n} = \frac{1}{k^2_\Delta}\left(\frac{1+\rho_\Sigma}{\rho^2_\Sigma}\right)\frac{B_\Delta}{B_{IF}}\frac{T_\Delta}{T_\Sigma} \approx \frac{1}{k^2_\Delta\, \rho_\Sigma}\left(\frac{B_\Delta\, T_\Delta}{B_{IF}\, T_\Sigma}\right) \qquad (8)$$

where ρ_Σ is the sum channel (IF) signal-to-noise ratio.

It should further be noted that the use of B_Δ as the two-sided noise bandwidth of the servo position loop facilitates the use of the above expressions as a measure of the total angle-tracking error due to thermal noise, i.e., the sum of two orthogonal tracking axes.

Finally, it may also be noted that the above expression is analogous to that given by Develet[(1)] for the case of equal system noise temperatures in the sum and difference channels (untypical in satellite communication terminals), and to that given by Manasse[(2)] when the signal detection time is equated to $1/B_\Delta$.

(1) J.A. Develet, "Thermal Noise Error in Simultaneous Lobing and Conical Scan Angle-Tracking Systems," IRE Trans. on Space Electronic and Telemetry, June 1961.

(2) R. Manasse, "Maximum Angular Accuracy of Tracking a Radio Star by Lobe Comparison," IRE Trans. on Antennas and Propagation, Jan. 1960.

APPENDIX B

MONOPULSE ERROR SIGNAL SLOPE

The following describes the selection of a "conditionally optimum" feed offset angle, or squint angle, which in turn produces an optimum monopulse tracking error signal slope in a classic amplitude comparison monopulse tracking system. The error slope parameter is considered as a basic figure of merit in monopulse tracking systems, and can be more completely defined as the slope of the monopulse difference (voltage) pattern at the boresight null, normalized by the sum channel signal voltage (via AGC).

In general, the slope of the difference pattern produced by a monopulse feed system is best determined experimentally; however, the approximation to follow can serve as initial system design criteria.

The main beam of the antenna may be represented by a voltage function of gaussian distribution:

$$E\ (\Theta) = Ae^{-\frac{p^2\ \Theta^2}{2}} \tag{1}$$

where A is an amplitude factor, p is an antenna geometry constant referred to as the beam shaping factor, and Θ is the angle off boresight. An analysis of an amplitude comparison monopulse system can be improvised by considering two such beams (corresponding to one axis), each displaced from the boresight by an equal but opposite amount, $\Delta\Theta$. The expressions for the two beams, A and B, are then, and

$$E_A(\Theta) = Ae^{-\frac{p^2(\Theta + \Delta\Theta)^2}{2}}$$

$$E_B(\Theta) = Be^{-\frac{p^2(\Theta - \Delta\Theta)^2}{2}} \qquad (2)$$

In the monopulse system, the sum (or reference) channel is given by the sum of $E_A(\Theta)$ and $E_B(\Theta)$, and the error channel is given by the difference between the two beams. The sum channel signal (voltage) is then

$$E_\Sigma(\Theta) = Ae^{-\frac{p^2(\Theta + \Delta\Theta)^2}{2}} + Be^{-\frac{p^2(\Theta + \Delta\Theta)^2}{2}} \qquad (3)$$

and the difference channel signal is

$$E_\Delta(\Theta) = Ae^{-\frac{p^2(\Theta + \Delta\Theta)^2}{2}} - Be^{-\frac{p^2(\Theta + \Delta\Theta)^2}{2}} \qquad (4)$$

These expressions represent the signals in the sum and difference channels as produced in the monopulse comparator. Their characteristic form is maintained throughout the hetrodyning operations down to the (last) receiver I.F. stage. At the I.F. point, however, the basic form of the signals is altered by the automatic gain control circuitry. The effect of the AGC is to normalize both the sum and the difference signal amplitudes with respect to the amplitude of the sum signal. The output of the sum channel becomes constant in time, ideally, and the output of the difference channel becomes the ratio $E_\Delta(\Theta)/F_\Sigma(\Theta)$. A basic angle detection scheme is shown in Figure 1 which illustrates this point.

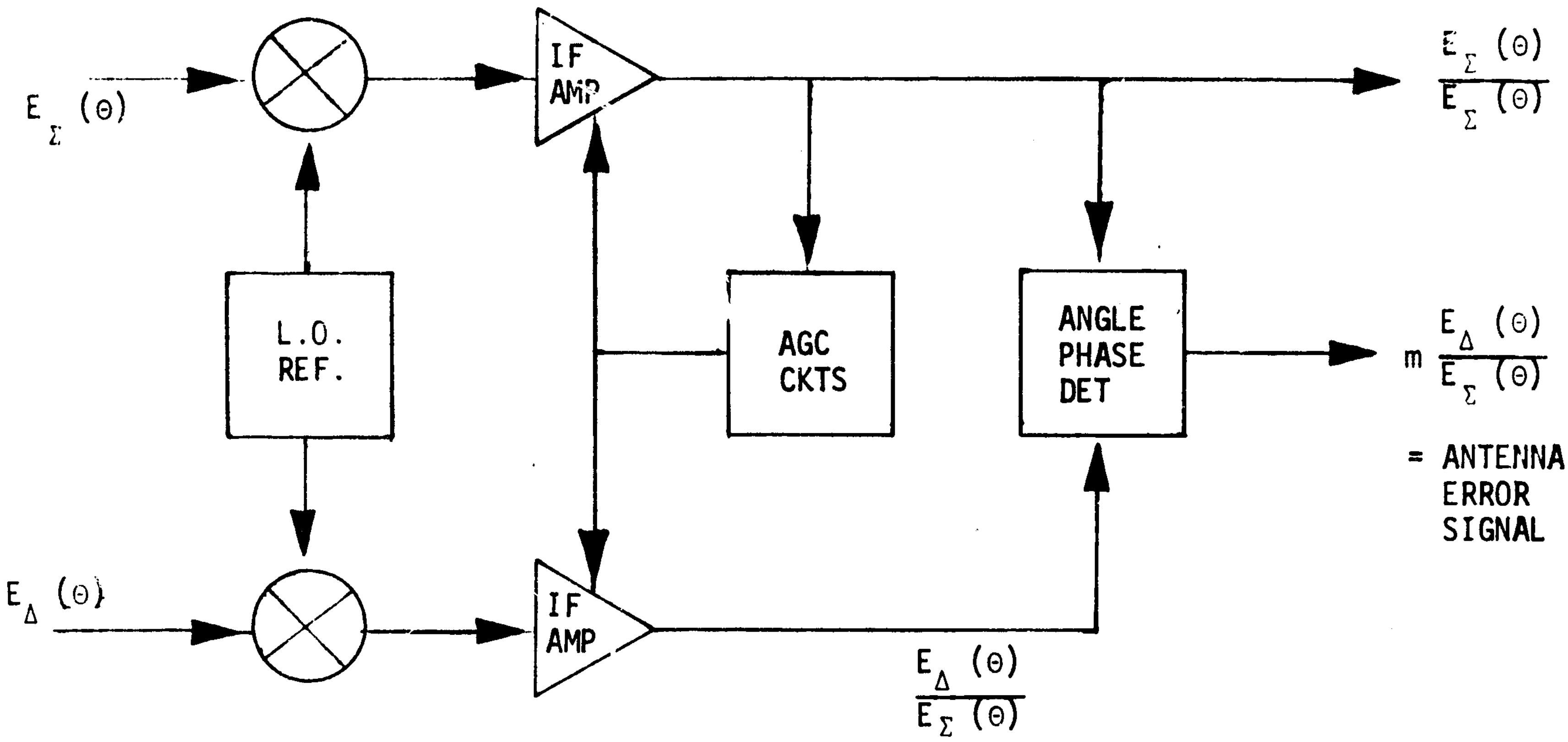
In a typical phase-locked angle-tracking receiver, the sum signal supplied to the angle phase detector may actually be the signal from the VCO or a reference oscillator employed in the sum channel. The resulting effect is, however, identical to that described above.

The effect of normalization via AGC is then such that the tracking sensitivity of the monopulse system is given by the slope of the signal (ratio) $E_\Delta(\Theta)/E_\Sigma(\Theta)$ near the boresight direction, i.e., as Θ approaches zero. This may be thought of as the slope of the signal that exists between the last difference channel I.F. amplifier stage and the (azimuth or elevation) angle phase detector, since the actual error signal at the output of the phase detector is modified by the phase detector gain constant, and the like, which does not in itself affect the basic monopulse tracking sensitivity. (Constants such as this are calibrated out.)

It may be assumed for purposes of this discussion that the basic beams given by Equation (2) have identical patterns, i.e., there is little if any amplitude unbalance, and thus $A \simeq B$. The normalized error signal is then given from Equations (3) and (4) by

$$\frac{E_\Delta(\Theta)}{E_\Sigma(\Theta)} = \frac{e^{-\frac{p^2(\Theta+\Delta\Theta)^2}{2}} - e^{-\frac{p^2(\Theta-\Delta\Theta)^2}{2}}}{e^{-\frac{p^2(\Theta+\Delta\Theta)^2}{2}} + e^{-\frac{p^2(\Theta-\Delta\Theta)^2}{2}}} \qquad (5)$$

As the tracking sensitivity, k_Δ, is defined as the slope of the signal above near the boresight direction, it may be determined



$$K_\Delta = \frac{\partial}{\partial\Theta}\left[\frac{E_\Delta(\Theta)}{E_\Sigma(\Theta)}\right]_{\Theta\to 0}$$

Figure B-1 BASIC ANGLE-DETECTION SCHEME

upon differentiation of Equation (5) with respect to the direction of signal source, Θ, and evaluating the result when the signal is on boresight, i.e., at $\Theta = 0$. It may also be noted that the slope of the ratio above at $\Theta = 0$ is numerically equivalent to the slope of the difference pattern divided by the sum channel gain at boresight. That is,

$$k_\Delta = \frac{\partial}{\partial_\Theta} \left[\frac{E_\Delta(\Theta)}{E_\Sigma(\Theta)}\right]_{\Theta \to 0} = \left[\frac{E_\Sigma(0)E'_\Delta(0) - E_\Delta(0)E'_\Sigma(0)}{\left[E_\Sigma(0)\right]^2}\right] = \frac{E'_\Delta(0)}{E_\Sigma(0)} \qquad (6)$$

since, by definition, the difference channel signal $E_\Delta(\Theta)$ is zero at $\Theta = 0$. In any case, the error slope at the boresight null is found, upon differentiation, to be

$$k_\Delta = -p^2 \Delta\Theta \qquad (7)$$

Thus, the theoretical value of the "normalized" slope of the monopulse difference pattern is given by the simple expression above comprised of the beam shaping factor "p" and the effective offset angle, or squint angle, $\Delta\Theta$. The beam shaping factor is essentially a constant for a given antenna, i.e., it is a function only of antenna size and frequency. It may be computed simply by evaluating the beam function, Equation (1), at the half-power point. The effective offset, or squint angle, on the other hand, is a fundamentally important parameter in tracking antenna system design.

For a given antenna and feed design, the effective offset angle for classic 4-horn feed geometry may be computed from

$$\Delta\Theta = \beta \tan \frac{d/2}{F} \approx \beta \frac{d}{2F} \qquad (8)$$

where d is the physical distance between a pair of feedhorns, F is the focal length of the antenna and β is the beam deviation factor. The beam deviation factor varies from 0 to 1, and is typically very near unity.(1)

The actual selection of the offset angle value in antenna system design is usually based on a compromise. It may be noted from Equation (7) that the larger the value of offset angle, the greater the error slope, or tracking sensitivity is. It might be expected, in general, that the offset angle which maximizes tracking sensitivity would be a solution of

$$\frac{\partial}{\partial(\Delta\Theta)} \left[\frac{E_{\Delta}'(0)}{E_{\Sigma}(0)}\right] = 0 \qquad (9)$$

however, no such practical solution exists, as may be noted from Equation (7) (as derived for a gaussian distribution). The same conclusion results if uniform or cosine amplitude distributions are assumed.(2) Another criterion that could be used is to maximize the slope of the difference pattern itself, i.e., employ the offset angle that results from

$$\frac{\partial}{\partial(\Delta\Theta)} \left[E_{\Delta}'(0)\right] = 0 \qquad (10)$$

In this case a solution does exist. As may be noted upon differentiation of the difference pattern of Equation (4), its slope is given by

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(1) Potter, P.D. "Antenna Design", Chapter 9 of "Space Communications" Ed. by Balakrishnan, A.V., McGraw-Hill, 1963.

(2) Rhodes, D.R., "Introduction to Monopulse". McGraw-Hill Book Co., 1959.

$$E'_{\Delta}(0) = -2Ap^2 \, \Delta\Theta \left[e^{-\frac{p^2 \Delta\Theta^2}{2}} \right] \qquad (11)$$

in which case the solution of Equation (10) results in

$$\Delta\Theta = \frac{1}{p} \qquad (12)$$

as the value of offset angle which maximizes the (un-normalized) slope of the difference pattern. This selection, however, generally results in a serious loss of sum channel gain. The sum channel gain on boresight is seen, by evaluating Equation (3) at $\Theta = 0$, to be

$$E_{\Sigma}(0) = 2A \, e^{-\frac{p^2 \Delta\Theta^2}{2}} \qquad (13)$$

This maximum theoretical gain of an antenna corresponds to a single feed on the boresight axis such that $\Delta\Theta = 0$. For a feed geometry with the effective offset angle given by Equation (12), the voltage pattern amplitude would be reduced by the factor $e^{-\frac{1}{2}}$, and the power pattern would be reduced by a factor of e^{-1}, corresponding to about 4.3 dB loss in reference channel gain. Clearly, this would not represent a good system design in a power-limited communications link. Thus, a trade-off exists between tracking sensitivity and sum channel gain in selecting the effective offset angle. As this compromising situation is typical, it has come about from experience that the optimum squint angle be taken as that which maximizes the product of sum channel amplitude and (un-normalized) difference pattern slope at boresight.[2][3] That is, for a classic four-horn monopulse feed system, the optimum offset angle

(2) Rhodes, D.R., Op. Cit.

(3) Hannan, P.W., "Optimum Feeds for all Three Modes of a Monopulse Antenna". IRE Trans. on Antennas and Propagation. Sept. 1961.

is that which results from

$$\frac{\partial}{\partial\,(\Delta\Theta)} \left[E_{\Sigma}(0) \;\; E_{\Delta}'(0) \right] \;=\; 0 \qquad (14)$$

Maximizing the product of Equations (11) and (13) then, by means of Equation (14), yields an "optimum" offset angle of

$$(\Delta\Theta)_{opt.} \;=\; \frac{1}{\sqrt{2}\,p} \qquad (15)$$

Employing this value for the effective offset angle results in a normalized monopulse error slope as given by Equation (7), with (15), of

$$(k_{\Delta})_{opt.} \;=\; \frac{-\,p}{\sqrt{2}} \qquad (16)$$

Further, the effect of selecting the latter value of offset angle on the sum channel gain at boresight may be found by substituting in Equation (13) with (15). The effect is to reduce the voltage pattern amplitude by a factor of $e^{-\frac{1}{4}}$, and thus the power pattern by a factor of $e^{-\frac{1}{2}}$. This corresponds to resultant antenna gain, referred to the sum channel, of about 2.17 dB less than the theoretical maximum permitted by a single (non-tracking) feed located on the boresight axis. This is analogous to limiting the maximum (reference channel) aperture efficiency to about 61%. (Factors other than feed offset would, of course, further reduce efficiency.) Thus, the selected value of offset angle and ensuing monopulse error signal slope may be considered as a practical "optimum" in system design criteria.

It is more convenient to express the optimum offset angle and error signal slope given by Equations (15) and (16) in terms of antenna half-power bandwidth, since this parameter is generally more readily available or approximated. Referring to the voltage function assumed in Equation (1), the basic <u>power</u> pattern is then of the form

$$p(\Theta) = A^2 e^{-p^2 \Theta^2} \tag{17}$$

Given the half-power beamwidth, the antenna beam slope factor "p" may be found from Equation (17) with $p(\Theta)/A^2 = \frac{1}{2}$, with $\Theta = \Theta_{hp/2}$, where Θ_{hp} is the two-sided half-power beamwidth. This is solved as

$$p = \frac{\left[-\ln \frac{p(\Theta)}{A^2}\right]^{\frac{1}{2}}}{\Theta} = \frac{2 (\ln 2)^{\frac{1}{2}}}{\Theta_{hp}} = \frac{1.67}{\Theta_{hp}} \tag{18}$$

Inserting this result into Equations (15) and (16), the final expressions for optimum offset angle and slope may be written as

$$(\Delta\Theta)_{opt.} = 0.424 \quad \Theta_{hp} \tag{19}$$

and

$$(K_\Delta)_{opt.} = \frac{1.17}{\Theta_{hp}} \tag{20}$$

Further, a final parameter of interest is the capture range, or capture beamwidth, of the angle tracking system. This is the angular offset over which the output difference channel voltage pattern, or "S" curve, gives a driving signal of the correct polarity. Clearly, it can be no greater than twice the feed offset angle (representing the peaks of the difference patterns). Thus, the capture range Φ may be approximated as

$$\Phi \lesssim 2\,\Delta\Theta = 0.85 \quad \Theta_{hp} \tag{21}$$

It may also be noted from the relationship of $p(\Theta)/A^2$ and Θ that 85% of the three dB beamwidth corresponds to the 2.2 dB beamwidth.

APPENDIX C

ANGLE TRACKING ERROR DUE TO THERMAL NOISE IN A SINGLE-CHANNEL MONOPULSE RECEIVER

Single channel monopulse is a hybrid angle tracking scheme which offers certain advantages inherent in both conventional multi-channel monopulse and conical-scan. It employs a monopulse feed rather than a mechanically rotating feed (or antenna) but yet requires only a single receiver channel. In addition to saving the cost of difference channels (preamplifiers, down-converters, IF's, etc.), the hybrid scheme negates the necessity of tedious periodic alignment of phase and gain between channels. These advantages are achieved at the expense of higher tracking errors and slightly reduced terminal net G/T performance.* This latter feature is due to the insertion of a coupler in the "sum" channel receive path prior to low-noise preamplification, which is used to add to (i.e., amplitude modulate) the sum channel a signal resulting from the combination, in phase or time, of the two difference channels.

Single-channel monopulse is alternately referred to as pseudo-conscan and monoscan. The following paragraphs review the calculation of angle tracking error due to receiver noise for a typical monoscan configuration. The simplified block diagram of Figure C-1 depicts the basic single-channel concept. The monoscan converter, which combines the two difference channels into a single multiplexed signal, may take a variety of forms based on either phase multiplexing (adding two signals in phase quadrature) or time-division-multiplexing

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* Antenna subsystem figure of merit for receive functions; the ratio of antenna gain to receiving system noise temperature.

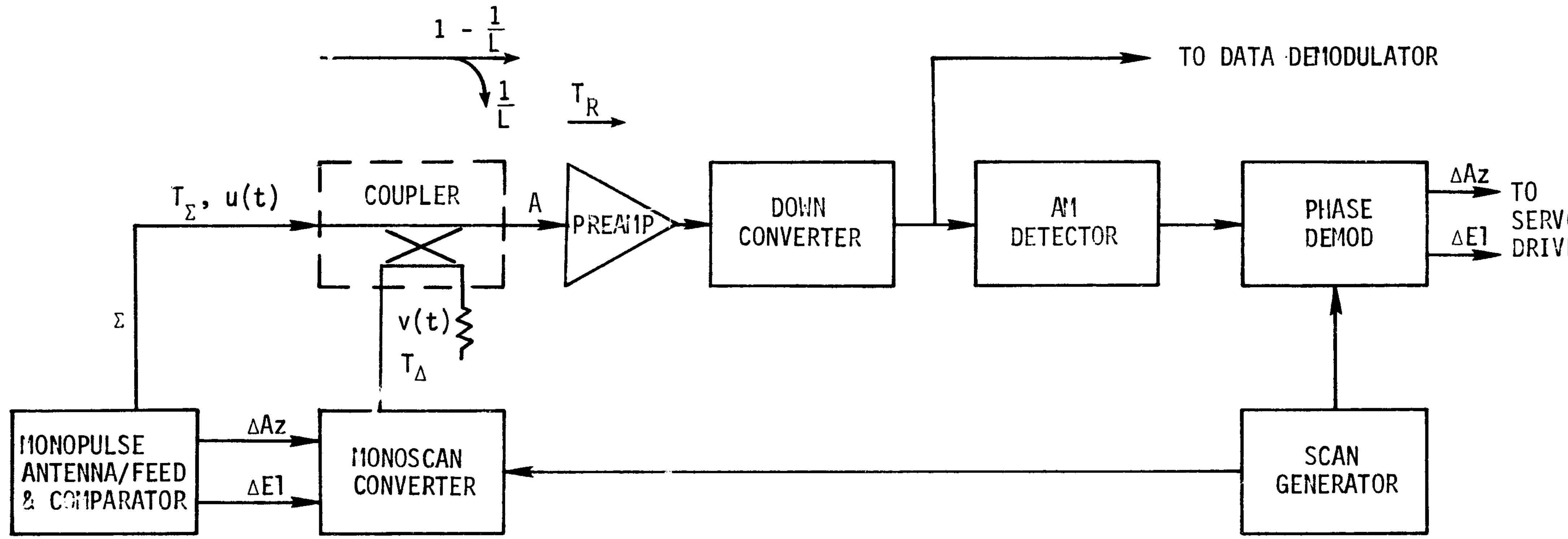

Figure C-1 SINGLE-CHANNEL MONOPULSE RECEIVER MODEL by alternate sampling. This analysis assumes the latter technique, which employs relatively simple digital control circuitry. The output of the difference channel multiplexer, the monoscan converter, is added to the sum channel via a simple coupler to produce a composite amplitude-modulated pseudo-conscan signal. After appropriate down-conversion to IF, the error signals are AM detected and the two tracking channels are separated by the phase demodulator (time-demultiplexer). The actual AM detector may be the communications receiver's AGC detector, and may be either a simple non-coherent envelope detector or a phase-locked coherent (quadrature) amplitude detector. The phase demodulator may consist of gated phase detectors.

The comparator sum channel output, u(t), may be expressed as follows:

$$u(t) = G(\Theta)\, S(t) + n_{\Sigma}(t) \tag{1}$$

where

$G(\Theta)$ = voltage gain of the main beam at an angle Θ off boresight

$S(t)$ = received signal at the antenna aperture

$n_{\Sigma}(t)$ = antenna sum channel noise

Then, the error channel, v(t), is given by:

$$v(t) = G(\Theta)\, k_{\Delta} \left(\frac{1}{L_C}\right)^{\frac{1}{2}} \left[U_A(t)\, \Theta_A + U_E(t)\, \Theta_E\right] S(t) + n_{\Delta}(t) \tag{2}$$

where

k_{Δ} = normalized (voltage) gain slope of the difference channel patterns

L_C = monoscan converter power loss

$n_{\Delta}(t)$ = error channel noise

$U_A(t)$ = unit azimuth channel sampling function

$U_E(t)$ = unit elevation channel sampling function

Θ_A = azimuth angle between the target and boresight

Θ_E = elevation angle between the target and boresight

The sampling functions $U_A(t)$ and $U_E(t)$ are orthogonal to each other as shown in Figure C-2

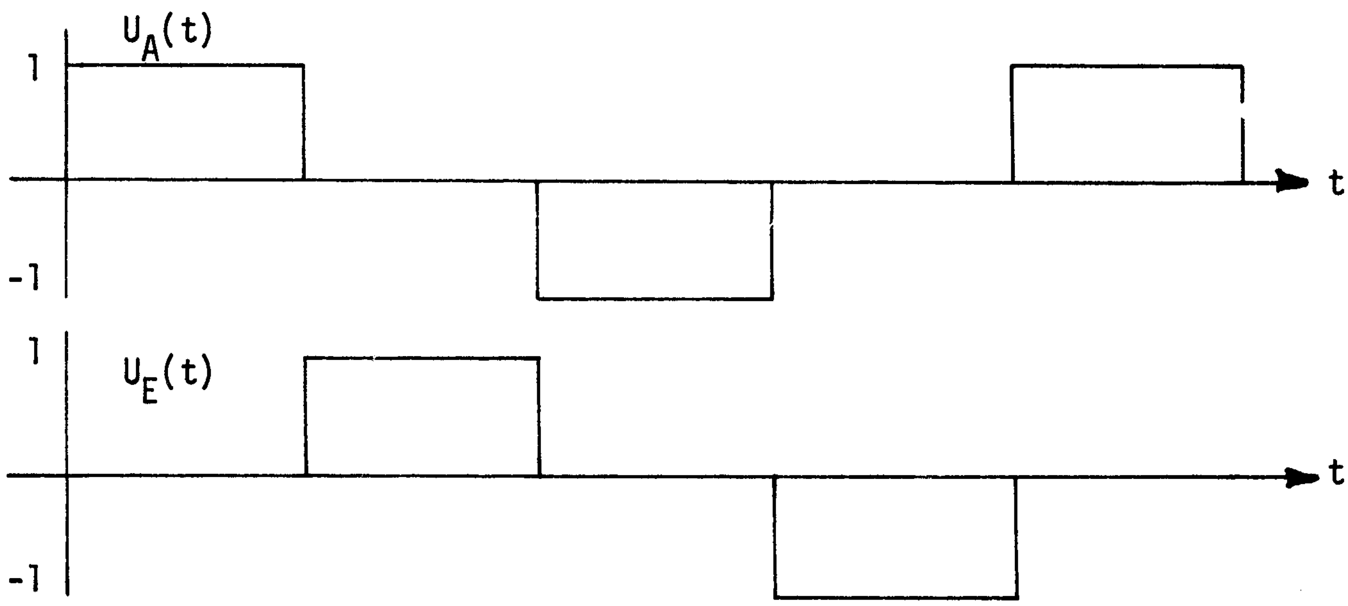

Figure C-2 Monoscan Sampling Function

The input to the preamplifier at Point A in Figure C-1, denoted by w(t), can be represented by:

$$w(t) = \left(1 - \frac{1}{L}\right)^{\frac{1}{2}} G(\Theta) \left\{1 + \left[\frac{1}{L_c\ (L-1)}\right]^{\frac{1}{2}} k_\Delta \left[U_A(t)\ \Theta_A + U_E(t)\ \Theta_E\right]\right\} S(t)$$

$$+ \left(1 - \frac{1}{L}\right)^{\frac{1}{2}} n_\Sigma(t) + \left(\frac{1}{L}\right)^{\frac{1}{2}} n_\Delta(t) + n_R(t) \qquad (3)$$

where $n_R(t)$ = receiver noise

L = coupling factor (as indicated in Figure C-1)

The first term of Equation (3) is the amplitude modulated signal, the amplitude of which is $(1 - 1/L)^{\frac{1}{2}} G(\Theta) \bar{S}$, where $\bar{S}$ is the amplitude of S(t). The modulation index k_m, is given by:

$$k_m^2 = k_\Delta^2 \frac{1}{L_c (L-1)} \tag{4}$$

This modulation index is a measure of the sensitivity of the tracking system to the target offset from boresight.

The system noise, $n_S(t)$, is given by the last three terms of Equation (3), i.e.:

$$n_S(t) = \left(1 - \frac{1}{L}\right)^{\frac{1}{2}} n_\Sigma(t) + \left(\frac{1}{L}\right)^{\frac{1}{2}} n_\Delta(t) + n_R(t) \tag{5}$$

Assuming that w(t) is automatic-gain-controlled before the AM detector and that the AGC bandwidth is much smaller than the pseudo-conscan frequency, the input to the AM detector is given by:

$$w'(t) = P \left\{1 + k_m [U_A(t)\, \Theta_A + U_E(t)\, \Theta_E]\right\} S'(t)$$

$$+ \frac{P}{\left(1 - \frac{1}{L}\right)^{\frac{1}{2}} G(\Theta)\, \bar{S}}\, n_s(t) \tag{6}$$

where P = constant output of the AGC amplifier

S'(t) = frequency translated received signal with unity amplitude

Assuming a square-law AM detector as a convenient model, the maximum low-frequency noise spectral density at the output of the detector is given by:

$$\Phi_{om} = 2\left\{\frac{p^4}{\left(1-\frac{1}{L}\right)^2 G^4(\Theta)\,\bar{S}^2}\,\Phi_{ns}^2\,B_{IF} + \frac{p^4}{\left(1-\frac{1}{L}\right) G^2(\Theta)\,\bar{S}^2}\,\Phi_{ns}\right\} \tag{7}$$

where Φ_{ns} = noise spectral density of $n_S(t)$
B_{IF} = IF bandwidth

After phase detection, the azimuth error signal spectrum is given by:

$$S_{az}(f) = \frac{p^4 k_m^2}{2}\,\Theta_A^2\,S_U(f) \tag{8}$$

where $S_U(f)$ = spectrum of $U(t)$

Since the servo subsystem is designed to compensate for an angle offset of Θ_A when the output of the AM detector has a spectrum as indicated in Equation (8) above, the rms tracking error due to noise is thus given by:

$$\sigma_{\Theta n}^2 = \frac{\Phi_{om}B_\Delta}{\frac{p^4 k_m^2}{2}} = \frac{4B_\Delta}{k_m^2}\left\{\frac{\Phi_{ns}^2 B_{IF}}{\left(1-\frac{1}{L}\right)^2 G^4(\Theta)\,\bar{S}^4} + \frac{\Phi_{ns}}{\left(1-\frac{1}{L}\right) G^2(\Theta)\,\bar{S}^2}\right\} \tag{9}$$

where B_Δ = two-sided servo noise bandwidth. Since $\Phi_{ns}B_{IF}$ is the noise power at Point A in Figure C-1 (but in the IF bandwidth) and $1/2\,(1 - 1/L)\,G^2(\Theta)\,\bar{S}^2$ is the signal power, Equation (9) may be reduced to:

$$\sigma^2_{\Theta n} = \frac{B_\Delta}{k_m^2 B_{IF}} \left\{ \frac{N^2}{S^2} + 2 \frac{N}{S} \right\}$$

$$= \frac{1}{k_m^2} \left(\frac{1 + 2 S/N}{(S/N)^2} \right) \frac{B_\Delta}{B_{IF}} \tag{10}$$

where S/N is the pre-detection signal-to-noise ratio in the IF bandwidth.

This result is similar to that given in Appendix A for conventional multi-channel monopulse. A major factor reducing the tracking accuracy available with a given feed and received signal-to-noise ratio is that due to the reduction in the effectiveness of the monopulse error signal slope k_Δ when transformed to a monoscan modulation index k_m. The amount of this reduction is dependent on the coupling factor L as given by Equation (4). A typical monoscan converter loss is about 1 dB, and a typical sum channel coupling factor is about 10 dB. Since tracking error $\sigma_{\Theta n}$ is inversely proportional to k_m and k_Δ, Equation (4) indicates an increase in error by a factor of about 3.4 (ignoring signal-to-noise ratio and system noise temperature factors). Decreasing the coupling factor clearly improves accuracy, however, it does so only at the expense of effective antenna gain. As indicated in Figure C-1, the coupler has a gain in the sum channel or communications path of 1 - 1/L, i.e., a line loss of

$$L_\Sigma = \frac{L}{L-1} \tag{11}$$

As this line loss exists prior to low-noise preamplification, it amounts to a loss in effective antenna gain. In addition, it serves to increase the noise temperature of the feeder line, which is an increase in system noise temperature. The noise temperature increase is given by:

$$\Delta T = \frac{L_\Sigma - 1}{L_\Sigma} T_o = \frac{T_o}{L} \qquad (12)$$

where T_o is the ambient temperature (nominally 288^o K). While this increase is actually in line temperature, it is approximately the same as the attendant increase in system noise temperature since the other factors (a decrease in the antenna temperature contribution to system noise and slightly higher actual coupler insertion loss due to resistive sources) are small and opposing. The general effect of decreasing coupling factor L on effective antenna gain and system noise temperature is indicated in Figure C-3.

Thus, the single-channel monopulse approach to angle tracking is characterized by both higher noise error and higher RF line loss, all of which reduce the operational G/T of the receive terminal. Nevertheless, this disadvantage is often cost effective when the advantages of receiver simplicity and reliability are considered.

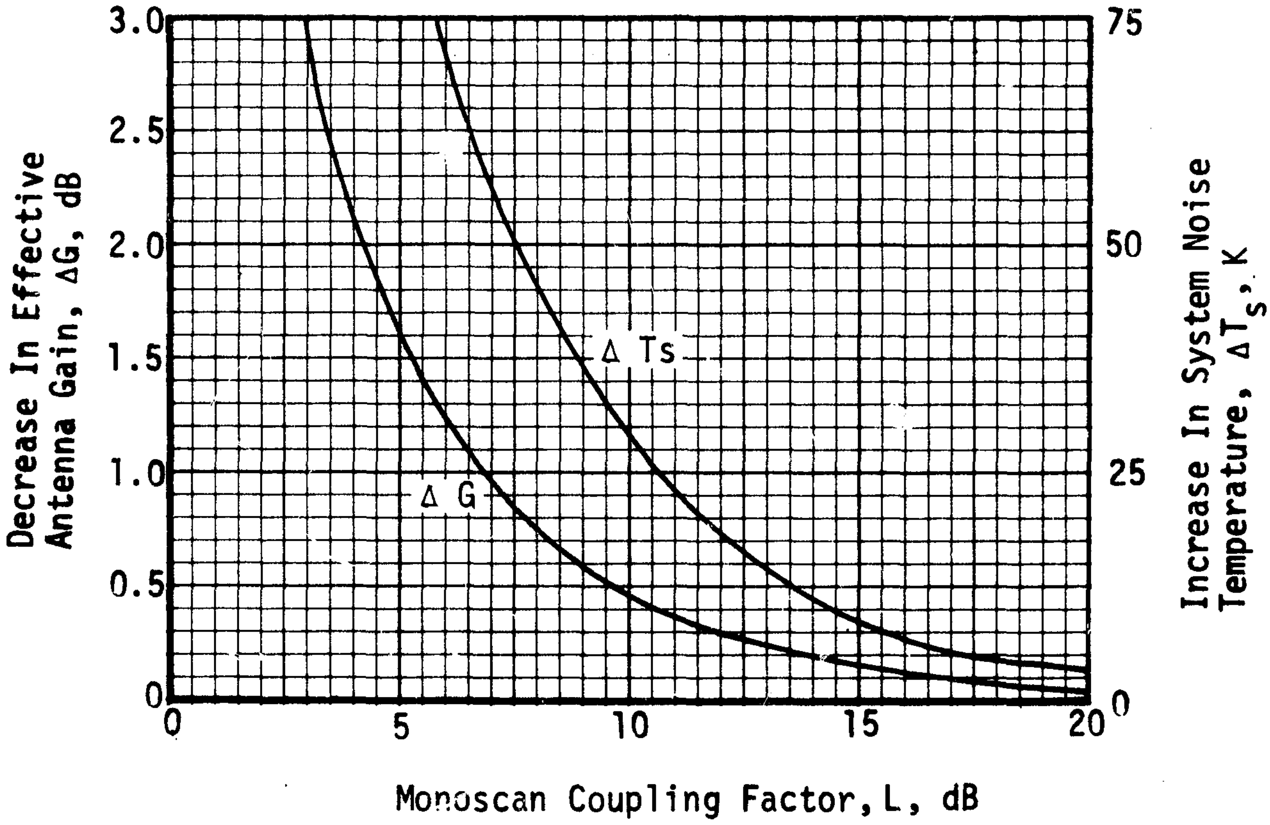

Figure C-3 EFFECT OF MONOSCAN COUPLING FACTOR ON TERMINAL RECEIVE G/T PERFORMANCE

APPENDIX D

PERFORMANCE ANALYSIS OF THE STEP-TRACK SATELLITE TRACKING TECHNIQUE WITH LOW GAIN SHIPBOARD ANTENNAS

1. GENERAL

Step-Track is a simple and inexpensive technique for automatic angle tracking of satellites with narrow beam antennas. The scheme is especially applicable to systems in which the relative angular dynamics between the satellite and earth terminals is small. No tracking feed or special RF receiver is required. After initial signal acquisition, which may be manual or remotely commandable, the antenna automatically makes a small angular move. By comparing the received signal levels before and after the move, the direction of the next move is decided. This process is continuous, alternating between antenna axes, such that the antenna is made to Step-Towards-Energy-Peak. The angular stepping is continuous, with step size selected so that the effect on operational antenna gain is small.

No special receiver is required to implement this autotracking function. The output of the communication receiver's AGC detector can be used as the received signal reference, with only simple sample-and-hold, timing and decision circuitry required to perform the Step-Track function. Of course, it is axiomatic that a tracking concept based on locating a beam maximum can never be as accurate as one which is based on locating a sharp pattern null, such as developed with conventional monopulse difference patterns. The scheme is also susceptible to AM interference at the stepping rate, similar to conical scanning tracking systems. However, the most significant limitation of the scheme is its unsuitability when the angular dynamics are high. Previous analyses[1] of the performance of the Step-Track technique was intended for relatively narrow-beam, fixed (ground-based) earth terminals operating with nominally synchronous-equatorial orbit communication satellites, where the angular dynamics is due only to very slow satellite drift.

It is the purpose here to extrapolate the referenced analysis for the case of wide-beam terminal antennas mounted on ships, in which ship's motion must be compensated.

2. APPROACH

Because the largest source of angular dynamics between a ship and a synchronous satellite is the roll component of ships motion, it is considered possible that an antenna with a fan beam could have advantages in minimizing required tracking rates. The pitch, yaw, heave, sway and surge angular excursions and rates are considerably smaller than that of roll about the ship's longitudinal axis. An antenna with a very wide elevation beamwidth may not need to track the roll motion, and a narrow beam in azimuth can be used to achieve the nominally required gain of about 10 dB. The azimuth angular rates caused by yaw and ships turning could be relatively slow and thus more easily compensated. While a scheme tailored for this concept would have distinct limitations when the satellite is in the bow or stern direction (along the longitudinal axis), for most

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(1) Tom, N.N. and Heckert, G.P., "Step-Track - A Simple Autotracking Scheme for Satellite Communication Terminals." Paper 70-416, AIAA 3rd Communications Satellite Systems Conference, Los Angeles, Calif. April 6-8, 1970.

angular positions the azimuthal rate requirements would be small.

In addition, the use of a fan beam as a model simplifies the analysis of ship-track performance since tracking need be accomplished in only one axis. As a baseline example, the following parameters are assumed:

Baseline Configurational Parameters

Antenna Beam Shape:	Fan Beam
Antenna Beamwidth in Azimuth:	Θ_{hp} (Az) = 20 deg
Antenna Beamwidth in Elevation:	Θ_{hp} (El) = 100 deg
Azimuthal Tracking Rate Required:	$\dot{A}$ = 10 deg/sec
Satellite Elevation Angle Range:	< Θ_{hp} (El)
Received Signal-to-Noise Ratio:	S/N = 10 dB in B_{IF}
IF pre-detection Bandwidth:	B_{IF} = 1 kHz

The fan beam defined by the above half-power beamwidths would provide for a peak antenna gain of about 12 dB. Tracking errors as high as one-half the 2 dB to 3 dB beamwidth would then be permitted. The azimuthal rate requirement of 10 deg/sec is quite high relative to yaw or turning; it is intended to accommodate the near-worst case effects of roll rate on azimuth angle when such geometric conditions exist (i.e., satellite in direction of bow-stern line). The assumed received carrier-to-noise density of 40 dB-Hz, corresponding nominally to a signal-to-noise ratio of 10 dB in a 1 kHz receiver IF bandwidth, is based on a near-worst case link situation and angle tracking on such a narrow band signal. This signal could be a special non-working channel designed only for system access control, and thus continuously transmitted by the satellite at all times.

In general, the tracking performance is found to be a function of the angular step size, the stepping frequency, the received signal-to-noise ratio, the signal bandwidth, the post-detection bandwidth and the relative angular rate between the satellite and terminal. The analysis contained herein follows the procedures developed in the referenced earlier work[1]. The tracking performance is obtained by first calculating the probability of a decision error after a step. Then, the stationary probability of the satellite being at any point is determined. Finally, from the stationary probabilities, the average loss in antenna gain is obtained. This probabilistic analysis requires characterization of the antenna gain function (assumed to be Gaussian), determination of the probability of error when comparing the two sequenced output voltage samples in a noisy environment (square-law detection assumed), derivation of the transitional probabilities between steps and the stationary probability of a given antenna pointing offset (a random walk), and calculation of the corresponding antenna gain loss as a function of angular step size $\Delta\Theta$, detector input S/N and pre-comparator (LPF) bandwidth b. The basic concept is illustrated in the simplified block diagram of Figure D-1.

3. EXAMPLE LOW GAIN ANTENNA PERFORMANCE

For a fan beam antenna whose beamwidth in the elevation direction is much wider than the maximum elevation drift of the satellite, it will only be necessary to step the antenna in the azimuth direction. The antenna may be stepped immediately after each sampling. Assuming that the satellite is moving in the positive azimuth direction, a typical satellite path with respect to the antenna azimuth boresight axis is shown in Figure D-2. As indicated in the figure, the effective satellite step size is $\Delta\Theta + \dot{\delta}T$, when the stepping is in the same direction as the satellite motion; and is $-\Delta\Theta + \dot{\delta}T$ when the antenna

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(1) Tom, N., and Heckert, G., Op. Cit.

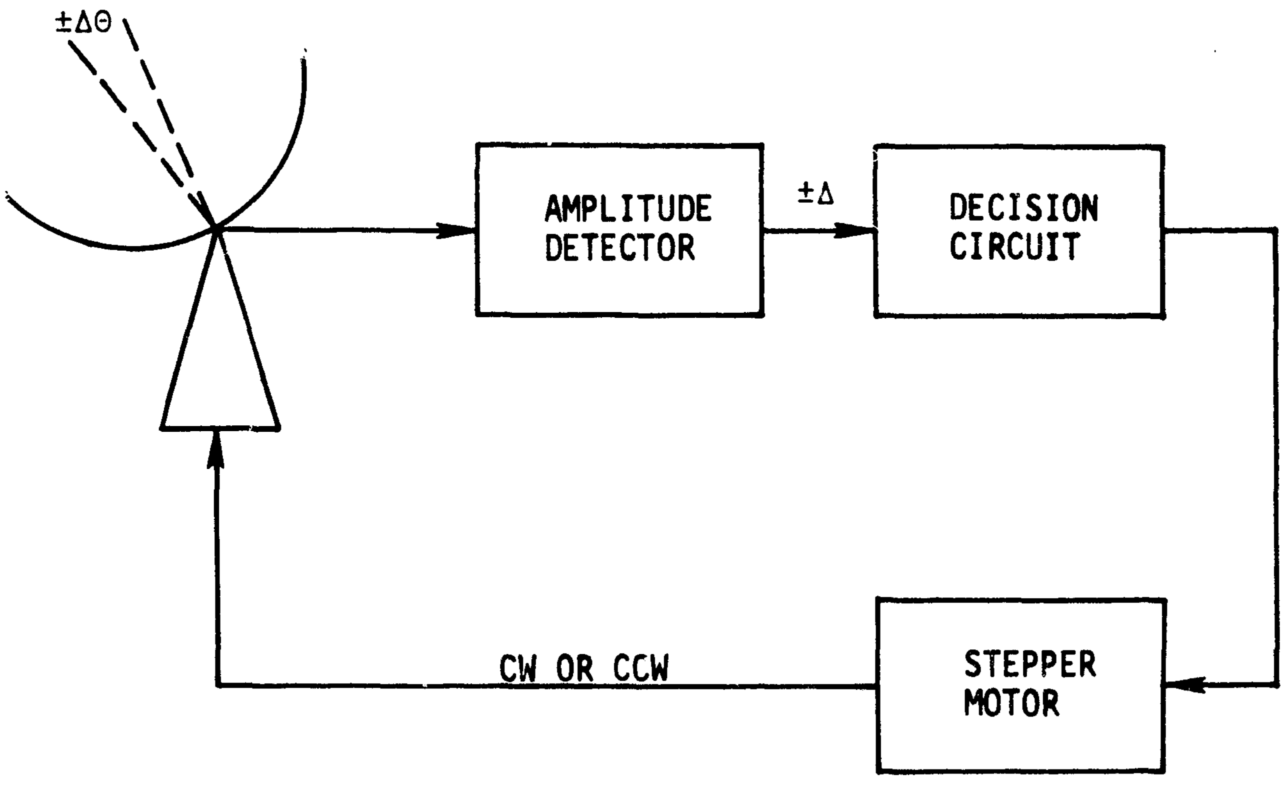

Figure D-1 SIMPLIFIED MODEL OF STEP-TRACK TRACKING LOOP

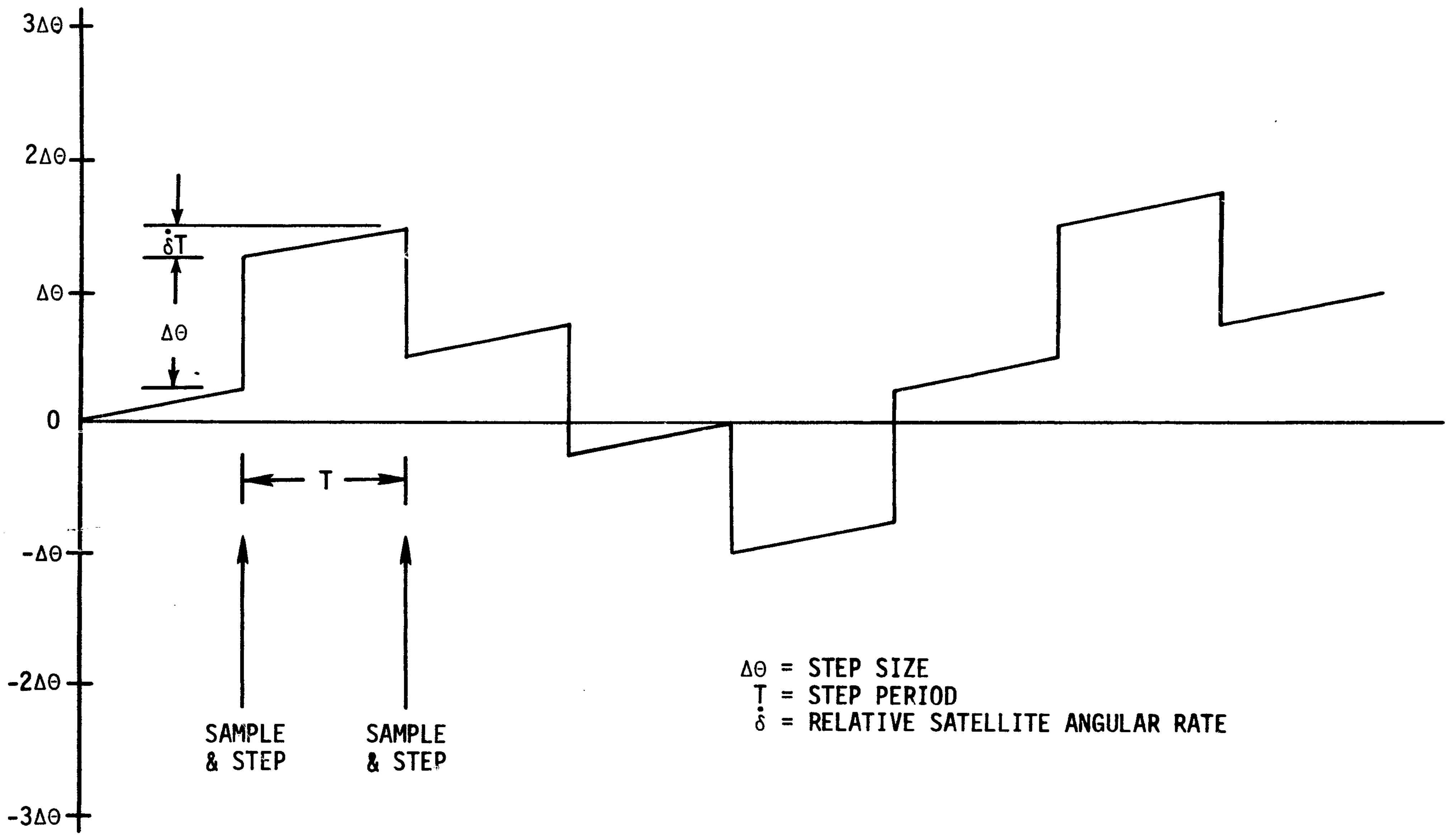

Figure D-2 RANDOM WALK PATH, ONE-AXIS TRACKING is stepped in the opposite direction. $\Delta\Theta$ is defined as the mechanical step size of the antenna; $\dot{\delta}$ is the satellite's relative angular rate; and T is the stepping time period. Obviously, $\dot{\delta}T$ must be smaller than $\Delta\Theta$ in order for the antenna to track the satellite.

As derived in Reference 1, the probability of a decision error after stepping from respective beam positions Θ_i to Θ_j is given by:

$$P_e\ (\Theta_i \rightarrow \Theta_j) = \frac{1}{\sqrt{\pi}} \int_{\Delta v_{ij}}^{\infty} e^{-x^2}\ dx \qquad (1)$$

where Δv_{ij} is the change in amplitude detector output dc voltage resulting from the angular step from Θ_i to Θ_j, and is given by:

$$\Delta v_{ij} = \frac{(S/N)\ \left| G(\Theta_i) - G(\Theta_j) \right|}{2\left(2\ \frac{b}{B_{IF}}\right)^{\frac{1}{2}} \left\{1 + \left(\frac{S}{N}\right) [G(\Theta_i) + G(\Theta_j)]\right\}^{\frac{1}{2}}} \qquad (2)$$

and where

- b = One-sided post-detection filter bandwidth
- B_{IF} = IF bandwidth
- $G(\Theta_i)$ = Normalized antenna power gain at an angle Θ_i off the boresight axis = $\exp\ [-2.77\ (\Theta_i/\Theta_{hp})^2]$
- S/N = Signal-to-noise ratio in B_{IF}

Based on the preceding discussion and on Figure D-2, it may be noted that Θ_j is related to Θ_i by:

$$\Theta_j = \begin{cases} \Theta_i + \Delta\Theta + \dot{\delta}T; & \text{for stepping in the same direction as the relative satellite motion,} \\ \Theta_i - \Delta\Theta + \dot{\delta}T; & \text{for stepping in the opposite direction as the relative satellite motion.} \end{cases}$$

The post-detection filter bandwidth is restricted by the fact that the detection circuit must wait until the output voltage of the post-detection filter settles to a steady-state level after each step before sampling. Since the rise time of a low-pass filter with a half-power bandwidth b is approximately equal to 0.5/b, and a reasonable time period for allowing the voltage to settle is about 3 times the rise time, the post-detection low-pass filter bandwidth should be selected to be about

$$b = \frac{1.5}{T} \tag{3}$$

Accordingly, Equation (2) may be resolved to the following:

a. For stepping in the same direction as the relative satellite motion,

$$\Delta v_{ij} = \frac{(S/N)\ |\ G(\Theta_i) - G(\Theta_i + \Delta\Theta + \dot{\delta}T)\ |}{2\left(\frac{3}{TB_{IF}}\right)^{\frac{1}{2}} \left\{1 + \left(\frac{S}{N}\right)[G(\Theta_i) + G(\Theta_i + \Delta\Theta + \dot{\delta}T)]\right\}^{\frac{1}{2}}} \tag{4}$$

b. For stepping in the opposite direction as the relative satellite motion,

$$\Delta v_{ij} = \frac{(S/N)\ |\ G(\Theta_i) - G(\Theta_i - \Delta\Theta + \dot{\delta}T)\ |}{2\left(\frac{3}{TB_{IF}}\right)^{\frac{1}{2}} \left\{1 + \left(\frac{S}{N}\right)[G(\Theta_i) + G(\Theta_i - \Delta\Theta + \dot{\delta}T)]\right\}^{\frac{1}{2}}} \tag{5}$$

From Equations (4) and (5), together with Equation (1), the probability of a decision error after stepping from any angle, Θ_i, to an adjacent angle Θ_j can be calculated as functions of the stepping period, T, and the mechanical angular step size $\Delta\Theta$.

Given the probability of a decision error, the transition probability (the probability of the relative satellite position changing from the discrete points Θ_i to Θ_j) can be calculated by the following equations:

$$P\,[\Theta_i \rightarrow (\Theta_i + \Delta\Theta + \dot{\delta}T)] =$$

a. $$P\,[(\Theta_i + \Delta\Theta - \dot{\delta}T) \rightarrow \Theta_i] \cdot P_e\,[(\Theta_i + \Delta\Theta - \dot{\delta}T) \rightarrow \Theta_i]$$
$$+ P\,[(\Theta_i - \Delta\Theta - \dot{\delta}T) \rightarrow \Theta_i] \cdot P_e\,[(\Theta_i - \Delta\Theta + \dot{\delta}T) \rightarrow \Theta_i]$$

$$\text{for} \quad \Theta_i > \frac{\Delta\Theta}{2} + \dot{\delta}T \tag{6a}$$

b. $$P\,[(\Theta_i + \Delta\Theta - \dot{\delta}T) \rightarrow \Theta_i] \; P_e\,[(\Theta_i + \Delta\Theta - \dot{\delta}T) \rightarrow \Theta_i]$$
$$+ P\,[(\Theta_i - \Delta\Theta - \dot{\delta}T) \rightarrow \Theta_i] \{1 - P_e\,[(\Theta_i - \Delta\Theta - \dot{\delta}T) \rightarrow \Theta_i]\}$$

$$\text{for} \quad -\frac{\Delta\Theta}{2} + \dot{\delta}T < \Theta_i < \frac{\Delta\Theta}{2} + \dot{\delta}T \tag{6b}$$

c. $$P\,[(\Theta_i + \Delta\Theta - \dot{\delta}T) \rightarrow \Theta_i] \{1 - P_e\,[(\Theta_i + \Delta\Theta - \dot{\delta}T) \rightarrow \Theta_i]\}$$
$$+ P\,[(\Theta_i - \Delta\Theta - \dot{\delta}T) \rightarrow \Theta_i] \{1 - P_e\,[(\Theta_i - \Delta\Theta - \dot{\delta}T) \rightarrow \Theta_i]\} \tag{6c}$$
$$\text{for} \quad \Theta_i < -\frac{\Delta\Theta}{2} + \dot{\delta}T$$

The stationary probability (the long term average of the time the satellite is at position Θ_i) is then given by:

$$P(\Theta_i) = P\,[(\Theta_i + \Delta\Theta - \dot{\delta}T) \rightarrow \Theta_i] + P\,[(\Theta_i - \Delta\Theta - \dot{\delta}T) \rightarrow \Theta_i] \tag{7}$$

Since the satellite must be at one angle at a given time, the sum of the stationary probabilities must equal to one. That is,

$$\sum_{\substack{\text{all} \\ \text{possible} \\ \Theta_i}} P(\Theta_i) = 1 \qquad (8)$$

Finally, the average loss can be calculated by:

$$\bar{L} = 10 \text{ Log } [\sum_{\substack{\text{all} \\ \text{possible} \\ \Theta_i}} P(\Theta_i) \; G(\Theta_i)] \text{ dB} \qquad (9)$$

For a fan beam antenna with the half-power beamwidth in azimuth of 20 degrees, the azimuth rate of 10 deg/sec, and S/N of 10 dB in 1 kHz IF bandwidth, the stationary probabilities and the average loss are calculated for the following selected parameters:

step size $\Delta\Theta$ = 2 degrees
step period T = 0.1 second.

Based on the previous analysis,[1] a step size of one-tenth of the half-power beamwidth was found to be optimum. Thus, a step size of 2 degrees was selected for the subject example. Also, certain experiments have indicated that the stepping interval should be selected such that the relative satellite position not be allowed to drift through more than one-half of the step size between intervals of stepping of the antenna.

For the step size $\Delta\Theta$ = 2 degrees, this means the step period must be such that the relative angular motion of the satellite does not exceed one degree during the period. For the assumed maximum

(1) Tom, N., and Heckert, G., Op. Cit.

azimuthal rate of 10 deg/sec, this means that the stepping interval should be 0.1 second.

The stationary probability (Equation 7) for the baseline example is shown in Figure D-3 as a function of angular offset relative to the peak of the antenna beam. It may be noted that the most probable position is about 2.5 degrees away from the beam peak. This represents the probability of a certain amount of lag. If the ship were rolling back and forth by a small amount, this average lag would be reduced to zero. The model assumed in the analysis, however, is for a large roll excursion at 10 deg/sec, i.e., essentially a continuous roll. The average loss in operational antenna gain was calculated (Equation 9) to be

$$\bar{L} = 1.3 \text{ dB}$$

This result constitutes a long term average degradation in antenna gain relative to its peak value, and assumes that tracking is always maintained, i.e. that the system does not lose lock.

Actually, because of the high angular rate assumed, there would be a high probability of losing lock with the system. If an instantaneous error were such that no signal is received, i.e., the receiver AGC loses lock, then the system would probably not recover and manual re-acquisition would be required. Complete loss of signal can obviously be expected when the satellite is outside the antenna beam, e.g., at angle offsets corresponding to twice the half-power beamwidth. For the example beamwidth (Θ_{hp}) of 20°, the "loss of track" error angle can be taken as ±20°. The probability of losing track is plotted in Figure D-4, as a function of the number of steps taken.

This figure shows that the probability of losing lock is one per-

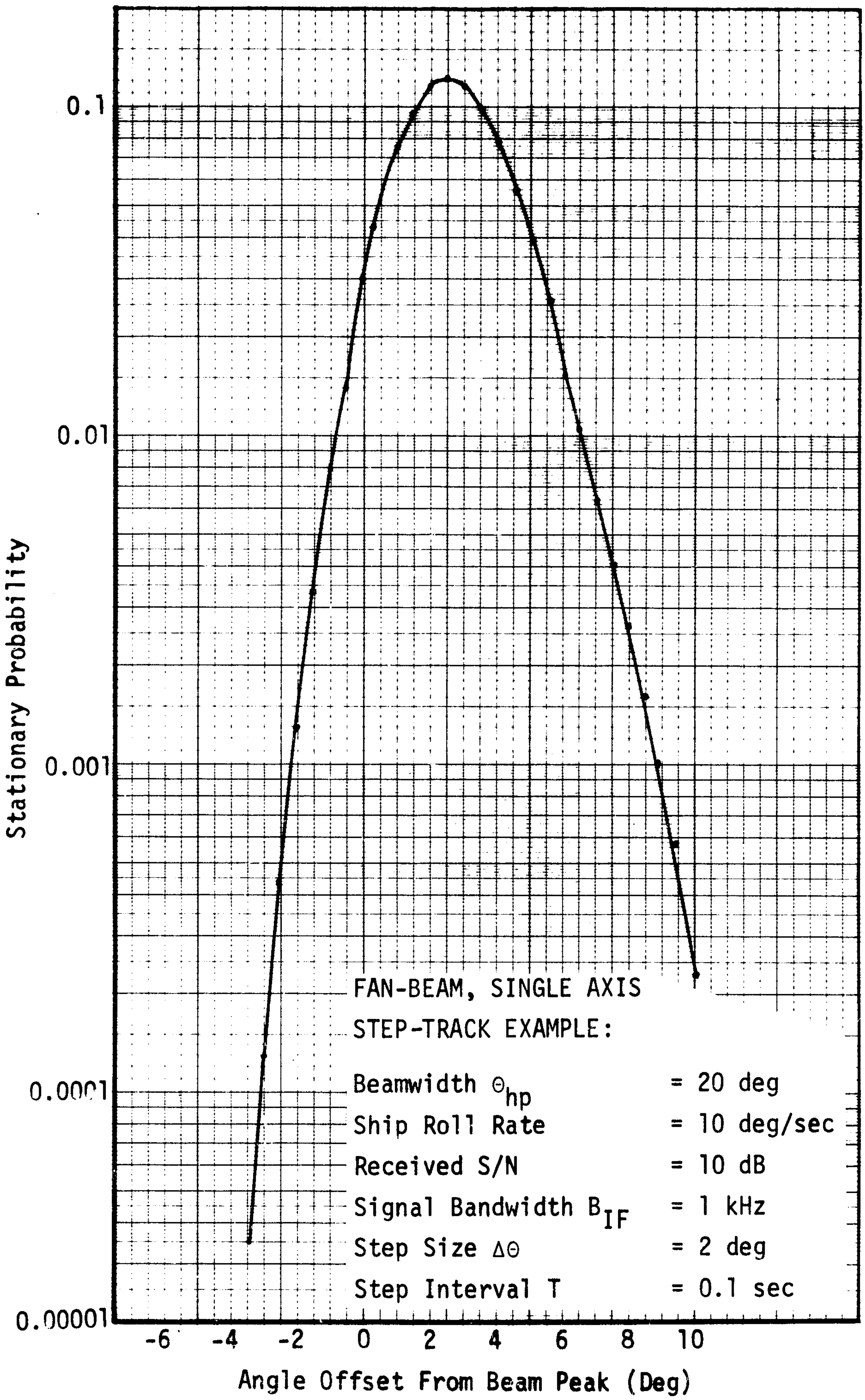

Figure D-3 STATIONARY PROBABILITY

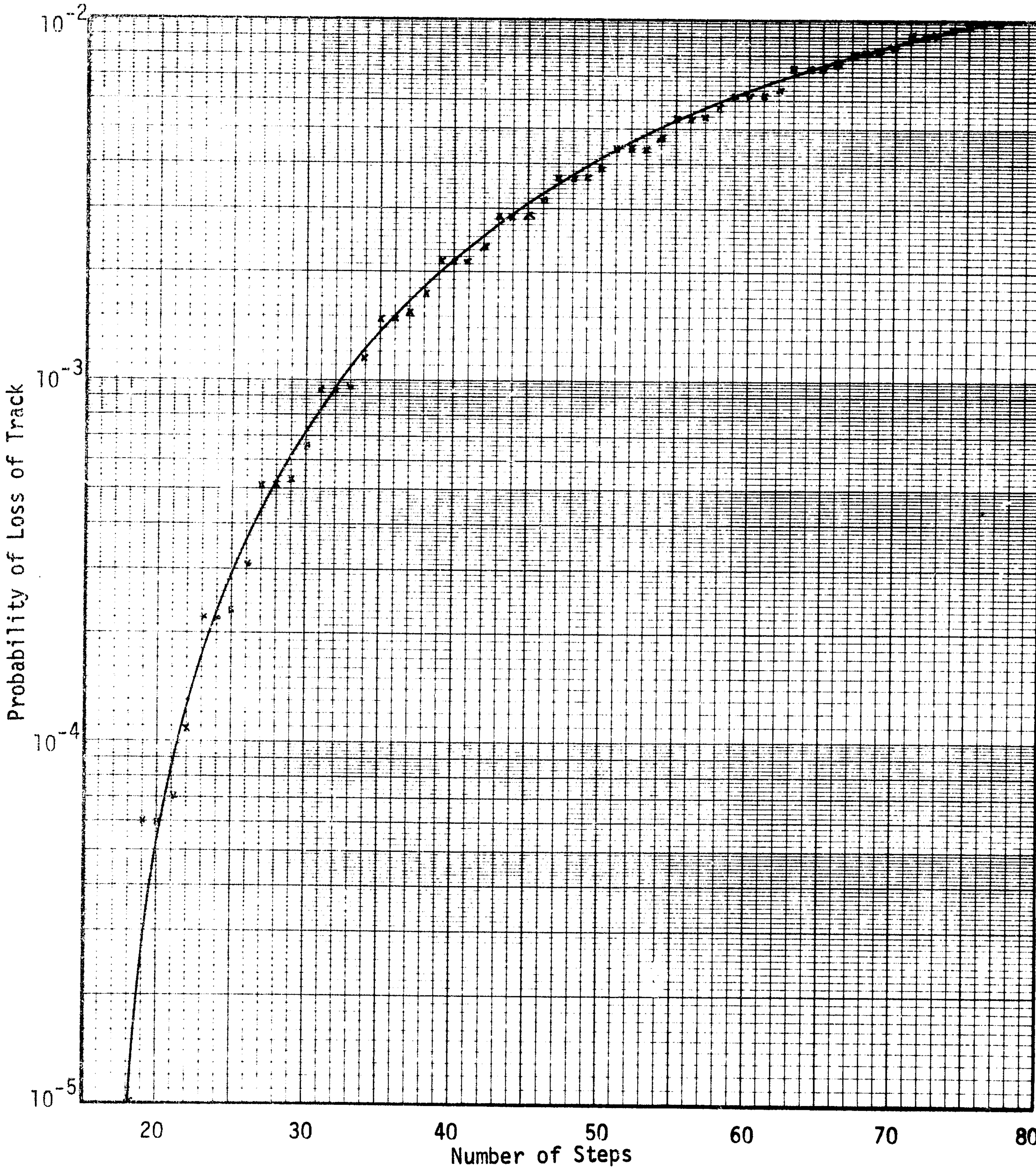

Figure D-4 PROBABILITY OF LOSS OF TRACK

cent when only 75 steps have been made. With the stepping interval of 0.1 second, this amounts to only 7.5 seconds. The probability of losing system lock could be as high as 10% in less than one minute.

On the basis of this analysis, it must be concluded therefore that the Step-Track technique would not be a credible tracking scheme for the example fan-beam antenna and assumed parameters of received C/N_0 (40 dB-Hz) and angular dynamics (10 deg/sec).

4. CONCLUSIONS

The preceding prediction of performance of Step-Track for single-axis tracking with a 20 degree beamwidth, with low level signals and high angular dynamics, has clearly indicated the unsuitability of the technique for shipboard use under such conditions. This conclusion could change significantly however, for different conditions. Pertinent conditional factors are the angular rates and received signal-to-noise densities.

Lower roll dynamics would obviously improve the performance, but fairly high rates must be specified for the shipboard antenna - even if a fan beam is used. Full realization of the benefits of steerable fan in azimuth only are not possible since ship roll dynamics have a major effect on azimuth rates when the satellite is in the direction of the bow-stern line.

The received carrier-to-noise density specified can have a major influence on performance. A higher C/N_0 can be in terms of either actual signal-to-noise ratio or a wider IF bandwidth with the same S/N, or both. Because of the assumed square-law detection model, wider IF bandwidth is not quite as effective as higher S/N at low values of B_{IF}, but the relationship is close enough to make the above generalization. It may be noted that a small change in the value of Δv_{ij}, as the lower limit in integral of Equation (1), can produce a major change in the probability of error, $P_e(\Theta_i \to \Theta_j)$. A few dB increase in C/N_0 can produce a significant decrease in error probability, analogous to that evident in the familiar curves of bit-error-rate vs received C/N_0 (or S/N) for digital data modulation techniques.

The importance of received C/N_0 as a tracking reference is not peculiar, of course, to the Step-Track technique; it applies as well to conventional (null-forming) angle tracking schemes. It is merely more critical at this particular level in Step-Track. In general, however, such factors serve to add motivation to enlarging the "size" of the access control channel. By this, it is meant that the access control channel could in general be at a higher data rate, by multiplexing data services (telex, etc.) together with the access control signal. The combined digital signal channel would have to be transmitted at all times of course, at full power. (The continuous transmission feature is the basic reason the fixed-frequency access control channel is considered most suitable as a tracking beacon.) A design value of C/N_0 of 45-46 dB-Hz would be most desirable in terms of ease of satellite transponder design, in that the control/data channel could then be allocated the same transmitter power as the voice channels.

It should also be noted that the preceding results for the fan beam example are generally applicable to certain other antennas, when attendant conditions are selected appropriately. Specifically, the curves of Figures D-3 and D-4 are applicable to a symmetrical antenna beam with same gain, i.e., an antenna with a beamwidth of 40-45 degrees in each axis. This antenna would have to be stepped twice as fast because of the requirement for two-axis stepping. The wider beam and higher rate tend to compensate for each other in terms of performance analysis.

One very important factor not heretofore noted is the implied rate requirements for the stepping drive of the antenna. In either the fan-beam or symmetrical-beam antenna examples, the required step period is 0.1 second, in terms of dwell time (per axis). The actual step through $\Delta\Theta$ should be accomplished in at least one-tenth this time, i.e., 0.01 seconds. With a two-degree step size, this implies an <u>average</u> drive speed of 200 deg/sec. Clearly, the associated accelerations and peak speeds required to meet the above average speed in 0.01 seconds would be a major imposition on an otherwise simple system concept.

Thus, even with higher received carrier-to-noise densities potentially available, it would appear that the Step-Track technique is not advisable for shipboard antennas, even those with wide beams.

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